A remote biological data acquisition system

by

Jonathan William Bredow

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LIST OF SYMBOLS AND ABBREVIATIONS

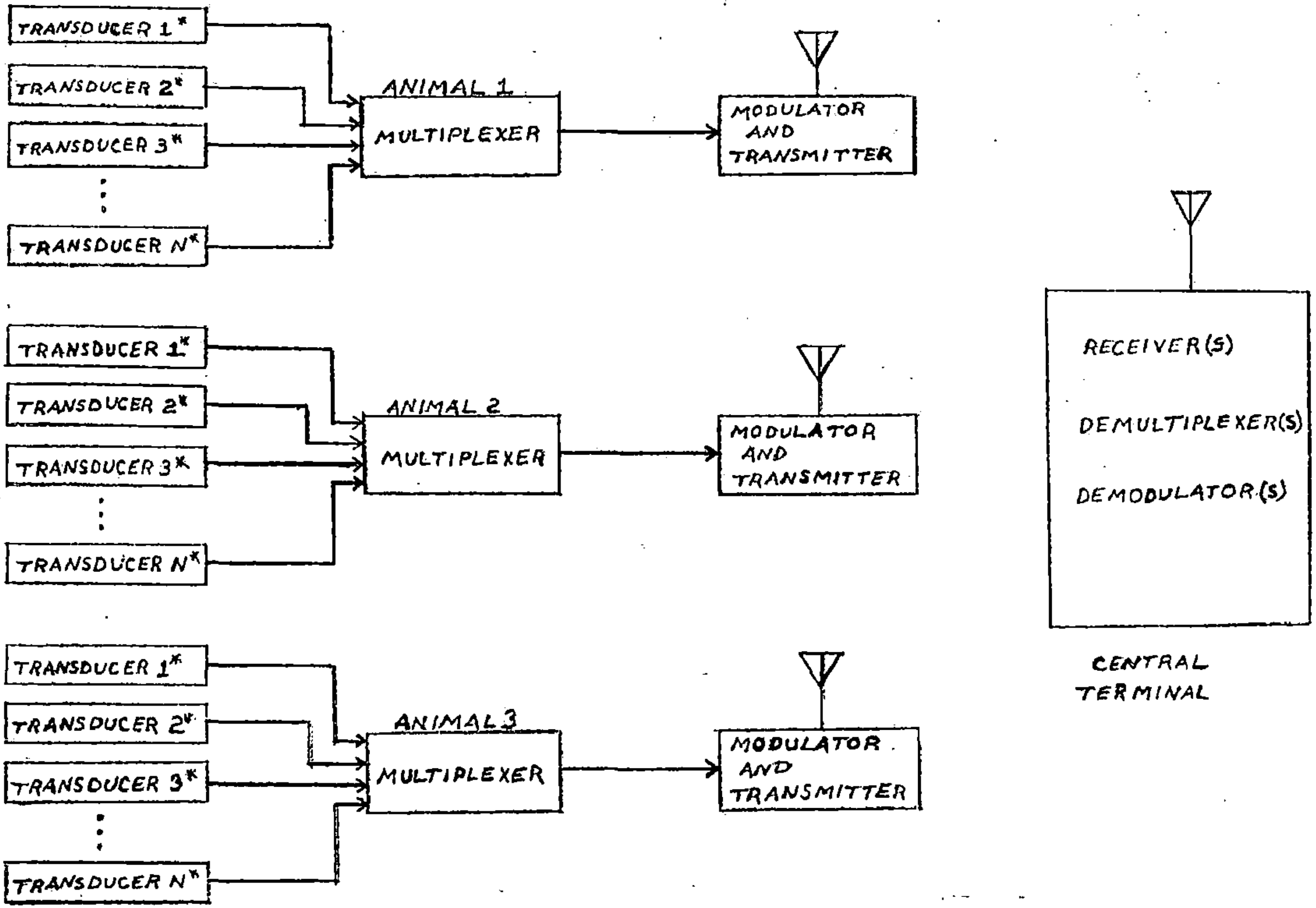
A	Ampere
AM	amplitude modulation
amp	amplifier
CB	citizens' band
CMOS	complementary metal-oxide semiconductor
db	decibel
DC	direct current
ECG	electrocardiagram
EEG	electroencephalogram
EPROM	erasable-programmable read-only memory
FM	frequency modulation
FSK	frequency-shift keying
GI	gastro-intestinal
H	Henry
hr	hour
Hz	Hertz
IC	integrated circuit
IF	intermediate frequency
JFET	junction field-effect transistor
k	kilo- or 10^3 (may be used alone or to form kA, kHz, kV, kW, kΩ)
m	milli- or 10^{-3} (commonly used to form mA, ms, mV, mW)
MHz	MegaHertz or 10^6 Hertz
NADC	National Animal Disease Center
Nicad	nickel-cadmium

op amp	operational amplifier
osc	oscillator
pf,pF	picoFarad or 10^{-12} Farads
PLL	phase-locked loop
PM	phase modulation
RAM	random-access memory
RF	radio frequency
s	second
UHF	ultra-high frequency
V	Volt
VCO	voltage-controlled oscillator
VFO	variable frequency oscillator
VHF	very high frequency
W	Watt
μ	micro- or 10^{-6} (commonly used to form μA, μF, μH, μV, μW)
Ω	Ohms

I. INTRODUCTION

Many types of biological data, i.e., ECG, blood pressure, EEG, etc., can easily be obtained in a laboratory environment. However, in many instances it is desirable to obtain such data from animals in their natural environment. A valuable method for obtaining such biological data from one or more free-roaming animals is to use radio telemetry. That is, each animal would carry, internally or externally, a radio frequency transmitter which would transmit a carrier frequency modulated with the biological signal(s) of interest. Of course, so that the signals of the transmitters do not interfere with one another, it would be necessary that each transmitter's carrier frequency be different. The biological signal(s) for each animal could then be recovered by a conveniently located receiver (containing the appropriate demodulation circuitry) tuned to the appropriate carrier frequency. In order to use only one receiver for several animals, some type of frequency scanning receiver might be useful. After recovery of the biological signals at the central terminal, these data could be stored or processed, as desired, by computer. A block diagram of such a data acquisition system is shown in Figure 1.

The concept of the biological telemetry system is simple enough, but the actual system components involved can be quite challenging to design and construct. This is especially true of the transmitters. For example, the transmitter should consume as little power as possible and be as small as possible. These two criteria are particularly important if the transmitter device is to be implanted. Other criteria which



* COULD ALSO BE ELECTRODES

Figure 1. Block diagram of a typical biotelemetry system

should be considered in the design and construction of a biotelemetry transmitter are:

1. Which type of modulation of the biological signal(s) onto the carrier frequency will produce the optimum combination of high signal/noise ratio at the receiver and low power consumption and small physical size at the transmitter?

2. How many biological signals will be transmitted from each animal?

   If more than one biological signal is to be transmitted using one transmitter, then either frequency or time multiplexing of the biological signals will be required. Thus, additional circuitry will be required to accomplish the task of multiplexing, which will in turn increase the size and power consumption of the transmitter device.

3. What type of material will be used to enclose the transmitter? If the device is mounted externally to the animal, then the circuitry and transducer must be kept from being exposed to the weather. If the device is to be implanted, swallowed, or otherwise used internally, then the transmitter circuitry and transducer must be protected from the harsh internal environment of the animal; more importantly, the coating of the device must be compatible with the homeostasis of the animal, so that toxicity does not result.

4. What type of transducer(s) will be used?

   Depending on the biological signal(s) to be obtained, variable capacitor, resistor and inductor transducers are the most common types. The design and construction of the transducers, however, is an art in itself.

5. What batteries are available?
   What battery sizes are available at the desired voltage, and what is the mA-hr rating of the battery? The answer to these questions will determine the life of the telemetry device once it has been placed on the animal. It may also have a significant bearing on the physical size of the telemetry package.
6. How much will the transmitted signal be attenuated if the device is implanted?
   Body tissues can greatly attenuate the transmitted signal; the amount of attenuation depends on the carrier frequency being used. Generally, the higher the carrier frequency is, the greater the signal is attenuated.
7. What carrier frequency is to be used?
   The components to be used in the transmitter will be dependent on the carrier frequency to be used; generally, the lower the carrier frequency is, the larger the physical size of the transmitter circuit components is.

These criteria apply to the transmitter device, but they would as equally well apply to a receiver which might be mounted on the animal. The purpose for such a receiver will become apparent in the following three chapters. A small portion of this paper discusses a remote device transmitter design, but the majority of the paper considers a number of different possible remote device receiver designs and more closely examines the one which is considered to be the most promising.

The remote device transmitter and receiver which are discussed in this paper are part of a universal biotelemetry device suggested by Professor Art Pohm.[1] This universal biotelemetry device features adaptability to various transducers, data filtering devices, amplifiers, etc., with a minimum of connections, and programmability (both central and remote), so as to allow a variety of data acquisition and animal experimentation routines to be used. This would enable not only biological data monitoring but also controlled injections, electric shocks, etc., during data monitoring. The control unit of the proposed unit would be a single chip CMOS microcomputer such as the 87C48 currently available from Intersil (15) which is a copy of the 8748 produced by Intel. The 87C48 contains 1 K bytes of EPROM and 64 bytes of RAM on the chip. The block diagram of this universal biotelemetry device is shown in Figure 3. This universal device will be further discussed in Chapter IV.

As can be seen in Figure 3, this universal biotelemetry device utilizes both a receiver and transmitter. This paper deals almost exclusively with the transmitter and receiver of the device and leaves the remainder of the unit design and construction to further research.

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[1] Art Pohm is a professor of electrical engineering at Iowa State University, Ames, Iowa.

II. BIOLOGICAL DATA TELEMETRY DEVICES CURRENTLY BEING USED

A. Remote Devices Utilizing Only a Transmitter

This discussion by no means includes every remote biotelemetry transmitter device that is now being used or that has ever been used but is an attempt to make comparisons of the characteristics of some of the more current devices in order that some generalizations about the characteristics of these devices can be made.

The first characteristic to be considered is the type of biological data being monitored. Table 1 shows characteristics of telemetry devices being used by different experimenters. Owen et al. (22) monitored the heart rates of birds. Slater and Bellet (29) monitored the body temperatures of submerged divers. Barr (2) used a telemetry device to monitor the body temperatures of large numbers of caged rodents. Brach et al. (4) monitored winterhardiness and temperatures of living plants by using telemetry. Klein and Davis (17) used biotelemetry to monitor EEG, ECG, peripheral pulse, and arterial blood pressure of surgical patients. Deutsch (8) presented a 15-electrode totally implanted biotelemetry unit for monitoring either various areas of the brain simultaneously or more localized areas in depth. Duncan et al. (9) used biotelemetry to measure the temperatures of avian shanks. A Technical Note in Medical and Biological Engineering (30) presented a telemetry device for remote monitoring of breathing pressures inside of a face mask under different exercise conditions. Clark and Dowdell (5) used telemetry to monitor pH and fluoride-ion concentration in dental plaque and saliva. Besseling et al. (3) used telemetry to monitor the electrical control activity of smooth muscle of the GI tract of unrestrained conscious animals. As can be seen, the possible uses for biotelemetry are almost unlimited.

The second characteristic of the devices in Table 1 to be compared is the carrier frequency of the transmitter. Of the telemetry devices listed in Table 1, Owen et al. (22), Klein and Davis (17), and Besseling et al. (3) used carrier frequencies within the FM radio band (88-108 Mhz). Of course, one of the primary reasons for using this range of frequencies is so that the receiving device can be constructed from an FM band receiver. Slater and Bellet (29) used ultrasound as the carrier frequency of their telemetry device because ultrasound transmission through water is more efficient than radio frequency transmission through water. Brach et al. (4), Ivison and Robinson (16), and Clark and Dowdell (5) also used VHF carrier frequencies. Duncan et al. (9), and the experiments reported in the Technical Note (30) used 27 MHz as the carrier frequency, probably because this frequency is in the citizen's band and also because a CB receiver can be used as part of the receiving device. Deutsch (8) indicated that for his purpose, 20 MHz was about the most ideal carrier frequency. Barr (2) used the AM radio band (530-1600 kHz) for his device. Some considerations which should be used for a given telemetry application include:

1. Is the remote device to be implanted?
   Higher frequencies tend to be more highly attenuated by living tissue than do lower frequencies, but circuits operating at higher frequencies tend to have smaller components.
2. What type of central terminal receiver is to be used?

Table 1. Characteristics of some current biotelemetry transmitters

INVESTIGATOR	OWEN ET AL. (22)	SLATER AND BELLET (29)	BARR (2)	BRACH ET AL. (4)	IVISON AND ROBINSON (16)	KLEIN AND DAVIS (17)	DEUTSCH (8)	DUNCAN ET AL. (9)	TECHNICAL NOTE (30)	CLARK AND DOWDELL (5)	BESSELING ET AL. (3)
CARRIER FREQUENCY	105-125 MHz	ULTRA-SOUND	590-1600 MHz	56 MHz (CRYSTAL)	VHF (CRYSTAL)	88-108 MHz	20 MHz	27 MHz	27 MHz	?	100-104 MHz
TYPE OF CARRIER MODULATION	FM	FM (DATA DIGITALLY ENCODED)	AM	FM (DATA DIGITALLY ENCODED)	PM	FM	FM	FM	AM	FM	FM
SUBCARRIER FREQUENCY	NONE	NONE	NONE	NONE	NONE	2.2 KHz 3.5 KHz 5.0 KHz 7.5 KHz	3200 Hz ADDED TO EACH CHANNEL	2 KHz	1100-1200 Hz	NONE	NONE
TYPE OF SUBCARRIER MODULATION	NONE	NONE	NONE	NONE	NONE	FM	↓	PULSE RATIO MOD.	FM	NONE	NONE
TYPE OF SIGNAL MULTIPLEXING	NONE	NONE	NONE	TIME MULTIPLEX	NONE	4-CHANNEL FREQ. MULTIPLEX	16-CHANNEL TIME MULTIPLEX	NONE	NONE	NONE	6-CHANNEL TIME MULTIPLEX
BIOLOGICAL DATA BEING MONITORED	SURFACE ECG SIGNAL OF BIRDS	BODY TEMP. OF SUBMERGED DIVER	BODY TEMP. OF CAGED RODENTS	WINTER-HARDINESS AND TEMP. OF PLANTS	NOT GIVEN	1-ECG 2-EEG 3-PULSE 4-ART. BP	15-ELECTRODES HUMAN BRAIN SIGS.	AVIAN SHANK TEMP.	PRESSURE VARIATIONS INSIDE FACE-MASK	PH AND FLUORIDE-ION CONC.	ELECT. CONTROL OF GI TRACT
POWER CONSUMPTION	.3MA (LIFETIME APPROX. 70 DAYS)	40 MW FOR TEMP. PLUG-IN	11 μA	REL. HIGH	6.5 MA AT 5V (OSC. STAGE)	3 MA AT 4V.	APPROX. 3.5 MA	4 V. 120-225 μA	REL. HIGH	LITTLE INFO.	177 μA
MAIN COMPONENTS	RF OSC.	PG. 635 (29)	RF OSC.	(4)	PG. 109 (16)	(17)	4-CMOS I.C.s, 1 LM 308 I.C. RF OSC.	PG. 545 (9)	PG. 105 (30)	PG. 159 (5)	4-CMOS I.C.s, 2-μ-POWER OP-AMP I.Cs RF OSC.

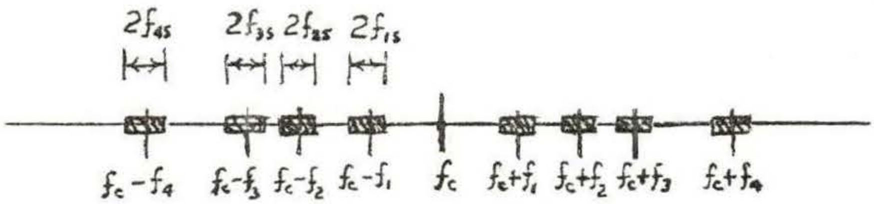
It is simplest as well as most economical to use a frequency that can be received by a commercially available receiver.

3. What frequencies will be clear of interference? The telemetry device should be designed to operate on a frequency for which no other commercial or private station can be received.

The third characteristic of currently used telemetry devices to be considered is the type of carrier modulation used. Of the devices listed in Table 1, eight experimenters used frequency modulation, two used amplitude modulation, and one used phase modulation. Ivison and Robinson (16) argued in favor of coherently detected phase modulation, stating that it satisfies the criteria of minimum transmitter power consumption, minimum bandwidth and operation over a short VHF radio link better than do the other types of modulation. However, frequency modulation has advantages over both amplitude and phase modulation. The biggest advantage that frequency modulation has over coherent detected phase modulation is that the FM receivers are simpler and commercially available. Frequency modulation has advantages over amplitude modulation in that it tends to be less distorted by noise than does amplitude modulation. Also, since signal strength at the receiver input tends to diminish as the transmitter is moved farther from the receiver, the data signal recovered from an AM signal, by the receiver, is dependent on the distance between the transmitter and receiver; on the other hand, data recovered from an FM signal, by the receiver, is independent of the distance between the transmitter and receiver since the signal frequency is unaffected by distance. With the integrated circuits which are now available, FM receivers need not be any more complex than AM receivers.

Some of the telemetry devices listed in Table 1 utilize subcarriers. The frequencies of the subcarriers used in these devices is the fourth characteristic to be considered. Of the devices listed in Table 1, only four of the experimenters used subcarriers. Three reasons for the use of subcarrier frequencies can be pointed out by using some of the devices in Table 1 as examples. Klein and Davis (17) used four subcarrier frequencies modulated onto the carrier frequency as a means of frequency multiplexing four channels of data. Figure 2 shows an illustration of frequency multiplexing for four channels of biological data (four biological signals) for an AM-AM system. A second reason for the use of a subcarrier is illustrated by Deutsch (8) when time multiplexing 16 channels of biological data onto the carrier. A 3200 Hz signal was added to each channel as a reference signal. A third reason for using a subcarrier is found in the device described in the Technical Note (30). This device employs an 1100-1200 Hz subcarrier which is frequency modulated by the pressure variation inside a face mask under different exercise conditions. This subcarrier is essentially unaffected by the surrounding environment, whereas the 27 MHz carrier will be much more greatly affected by the environmental conditions, since the changes in inductance and capacitance due to changes in environment will cause a much greater effect (percentage-wise) on the small inductance and capacitance values found in the 27 MHz circuitry than on the much larger inductance and capacitance values found in the 1100-1200 Hz circuitry. With the subcarrier, therefore, noise due to changes in environment is practically eliminated.

The type of subcarrier modulation used is the fifth characteristic of the devices in Table 1 to be considered. The reasons for using the



FREQUENCY SPECTRUM FOR 4-CHANNEL FREQUENCY MULTIPLEX AM-AM SYSTEM

f_{1s} IS MAXIMUM FREQUENCY OF SIGNAL MODULATED ONTO SUBCARRIER 1 (WHICH HAS A FREQUENCY OF f_1)

THE SAME REASONING FOLLOWS FOR THE OTHER 3 CHANNELS.

f_c IS THE TRANSMITTED CARRIER FREQUENCY.

NOTE: THE FOUR MODULATED SUBCARRIERS ARE SUMMED AND THEN MODULATED ONTO THE CARRIER.

Figure 2. Frequency spectrum of a 4-channel AM-AM frequency multiplexed biotelemetry transmitter

various types of subcarrier modulations are probably very similar to the reasons for using the various types of carrier modulation. Therefore, this will not be discussed except to mention that frequency modulation is again commonly used.

When more than one biological signal is being monitored, some type of multiplexing must be used. The type of multiplexing used is the sixth characteristic of the telemetry devices listed in Table 1 to be discussed. Of the four devices listed in Table 1 which monitor more than one biological signal, only one device utilizes frequency multiplexing; the others utilize time multiplexing. An example of an AM-AM frequency multiplexing scheme was shown in Figure 2. A time multiplexing scheme would not send a continuous signal of each of the biological signals being monitored. Instead, for an N-channel unit, the following time sequence would occur: First, the channel 1 data would be sampled and transmitted, then the channel 2 data would be sampled and transmitted. Then channel 3, channel 4, etc., and finally channel N. This sequence would occur continuously at a rate which would be several times the highest data frequency of any of the channels. Considering the components which are now available, it seems that time multiplexing would probably be simpler than frequency multiplexing due to the filtering that is required in demodulating a frequency multiplexed signal.

The seventh characteristic of the telemetry devices listed in Table 1 to be discussed is the power consumption. Unfortunately, the power consumption of the units is so widely varied that it is not possible to make general statements about devices. However, it can be seen that the simpler devices by Owen et al. (22), Barr (2), and Duncan et al. (9) consume very small amounts of power (< 1 mW). Surprisingly, the six-channel time multiplex device by Besseling et al. (3) draws only 177 μA at 4.5 v ($\sim$.8 mW). With the CMOS devices and micropower op amps which are now available, it is possible to build fairly complex low-power telemetry devices.

The final characteristic of the telemetry devices listed in Table 1 to be discussed is the amount of parts necessary for the telemetry device. Not all of the devices are considered, but only those which were the simplest single and multiple channel devices found. The other circuits can be compared by going to the references. The simplest telemetry devices are those by Owen et al. (22) and Barr (2). These devices consist of an RF oscillator as the main component. The 16-channel time multiplex device by Deutsch (8), however, requires 4 CMOS ICS, 1-LM308 op amp, and an RF oscillator as the main components. Likewise, the six-channel time multiplex device by Besseling et al. (3) uses 4 CMOS ICS, 2 micropower op amps, and an RF oscillator as the main components. These are not enough data to make generalizations, but it should be intuitively obvious that a multi-channel device requires more components than a simple single-channel device. The size of the multi-channel device will therefore probably be larger than a simple single-channel device.

These are the major comparisons of biotelemetry transmitter devices currently being used.

B. Remote Devices Utilizing Only a Receiver or Both a Receiver and a Transmitter

Descriptions of three recent remote data acquisition devices, which utilize only a receiver or both a transmitter and receiver, were found in the literature. One such device is presented by Cupal et al. (6) which utilizes a 570 kHz receiver as part of a biotelemetry repeater to be used on a large wild game animal. The repeater, which is mounted on the neck, receives biological data from an implanted 570 kHz transmitter and retransmits it on a VHF carrier to the central terminal. Another remote device which utilizes a receiver is presented by Guberek et al. (14). This receiver is part of a remote controlled coronary artery occluder which is used to study cardiovascular disease. Finally, Hudson in (10) presented a receiver to be used as a power switch for an implanted telemetry transmitter. When a radio frequency command is transmitted from the central terminal to this receiver, the receiver either supplies power to the implanted telemetry transmitter or removes it, depending on the command given. This receiver is discussed in more detail in Appendix A.

III. DISADVANTAGES OF CURRENTLY USED BIOTELEMETRY DEVICES

The biotelemetry devices discussed in Chapter II are useful for the applications for which they were constructed; however, these devices (with the exception of the device by Cupal et al. (6) which has characteristic 5) lack the following, sometimes desirable characteristics:

1. If the telemetry transmitter device moves out of range of the receiver or if the telemetry transmitter fails, the device on the animal should be capable of storing biological data long enough for the broken radio link to be restored.
2. The telemetry device on the animal should be able to do simple statistical analysis on the data so that the transmitted data can be minimized or eliminated (so that only necessary data are transmitted).
3. The remote device should be remotely programmable so that the data monitoring and/or experimentation routines can be altered from the central terminal.
4. The circuit connections of the transducers to the remote device should be minimal so that a wide variety of numbers of biological data signals can be monitored with a minimum of wiring connections or disconnections.
5. Since transducers sometimes must be located within the body, it would be convenient if the transducer information could be transmitted through the skin by means of a radio link to a more powerful, externally mounted, transmitter which could relay the data to the central terminal.

6. All telemetry transmitters should be capable of operating on a single frequency in order to reduce system design requirements and amount of frequency spectrum required.

IV. A UNIVERSAL REMOTE BIOTELEMETRY DEVICE

Since the telemetry devices discussed in Chapters II and III lack the desirable characteristics listed in Chapter III, a universal telemetry device which incorporates a single chip CMOS microcomputer controller, such as the Intersil 87C48 (15), is proposed; this device would at least partially fulfill the list of desirable characteristics given in Chapter III. This device was mentioned briefly in Chapter I and is shown in block diagram form in Figure 3. No estimate has been made for the power consumption of this device. However, micropower op amps and other low-power devices (such as small signal VHF and UHF transistors) are available for its construction; this should allow the power consumption of the entire device to be tolerable if the unit is mounted externally to the animal.

This device could fulfill the list of desirable characteristics of Chapter III as follows:

1. Sixty-four bytes of RAM are located on the 87C48 chip. Depending on how much other RAM is available in the device, it is capable of storing some biological data should a breakdown of the radio link between the remote device and the central terminal occur.
2. Simple statistical analysis of biological data can be done within the device by proper programming of the 1 K EPROM and by utilizing the 64 bytes of scratch pad RAM in the 87C48. Thus, the amount of data to be transmitted can be minimized in many cases. For example, maybe only heart rates outside of certain preset limits need to be transmitted.

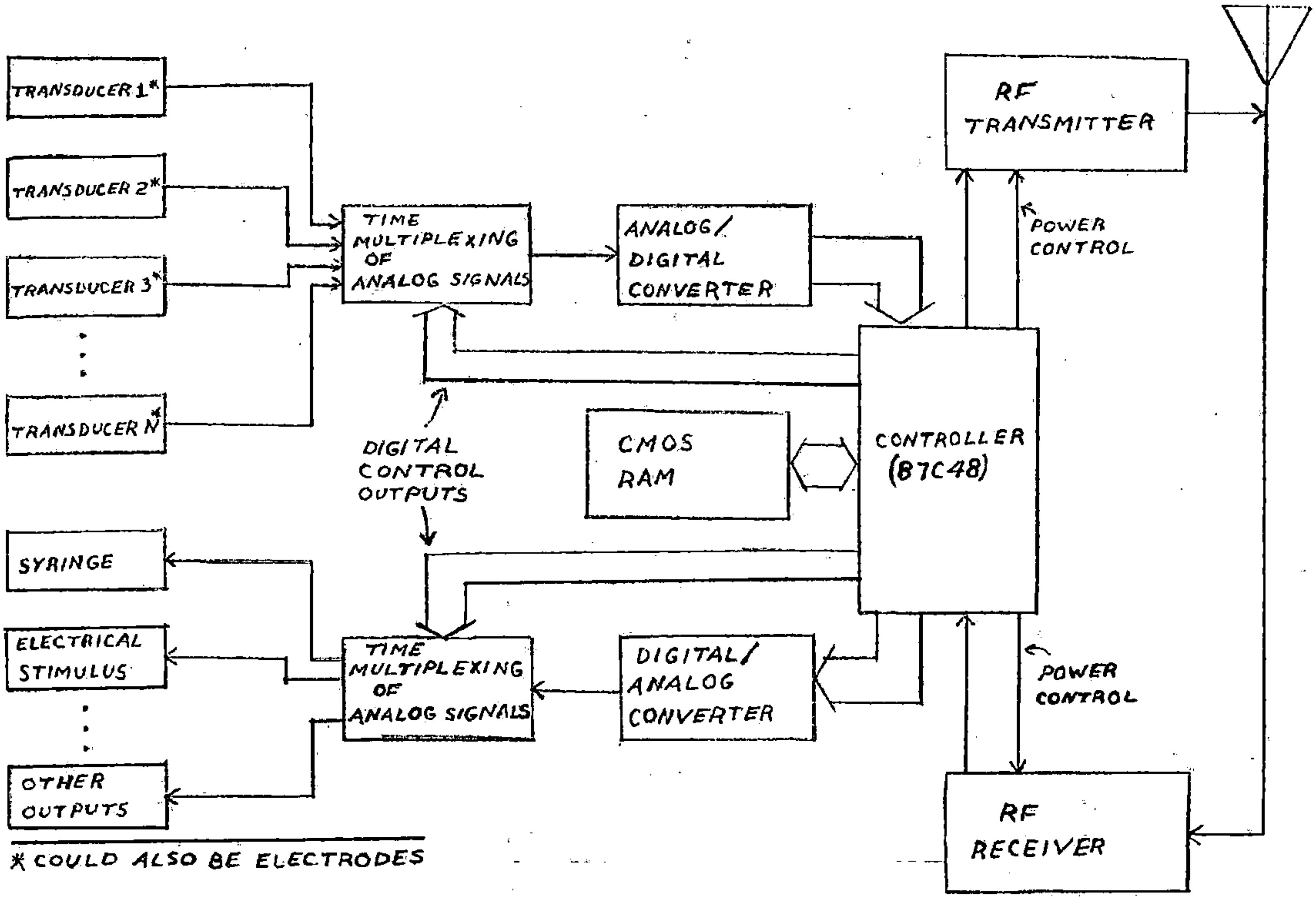


Figure 3. Block diagram of proposed remote data acquisition device

3. Again, depending on how much RAM is present in the remote device (in addition to the 64 bytes of RAM on the 87C48 chip), the device can be remotely programmed to carry out a variety of data monitoring and/or experimentation routines. Several routines could also be stored in the 1 K EPROM and called remotely.
4. It should only be necessary to connect or disconnect transducers or electrodes to the device in order to add or subtract biological signals to be monitored, since the multiplexing scheme can be altered by programming the 87C48 (minimize hardware alterations).
5. Since this device already incorporates a receiver, it should be relatively easy to program the remote device to receive and retransmit signals from a telemetry device located within the body.
6. Only a single carrier frequency is necessary for a biotelemetry system consisting of a large number of these devices since they can be addressed individually by the central terminal with a binary code.

As was mentioned in Chapter III, sometimes the above characteristics are not necessary; then it is better to use the simple telemetry devices listed in Chapter II, since a larger size and greater power consumption is necessary for the universal biotelemetry device. For many data acquisition systems, however, the versatility of the universal biotelemetry device and the other advantages that it has over the simple telemetry devices will outweigh the disadvantages. The remainder of this paper will concentrate on research done on the receiver and transmitter portions of this device.

V. EXPERIMENTAL PROCEDURE AND RESULTS

A. Remote Device Receiver Design

The experimental procedure which was followed throughout the receiver design research was divided into four parts. This was not a result of choice at the beginning of the project, but a result of other considerations which came into play during design and experimentation. The four parts of the experimental procedure are the AM approach, the pager approach, the superregenerative approach, and the weather radio approach.

Initially, it was decided that the superregenerative approach was inadequate due to the possibility of the units interfering if two animals carrying receivers tuned to the same frequency came close to one another. Thus, AM was initially considered to be the best approach in terms of power consumption and simplicity.

1. AM Approach

An amplitude-modulated signal consists simply of a carrier frequency, the amplitude of which is the data signal.

Briefly, an AM radio receiver (AM receiver is shown in block diagram form in Figure 4) functions as follows in recovering the data signal from a transmitted AM waveform.

The received signal is first amplified and then mixed with a local oscillator signal (a sine wave of frequency fo) such that signals of frequency fo + fc and fc - fo (fc is the received carrier frequency) are produced. A filter then blocks the signal of frequency fo + fc and passes the signal of frequency fc - fo. The signal of frequency fc - fo

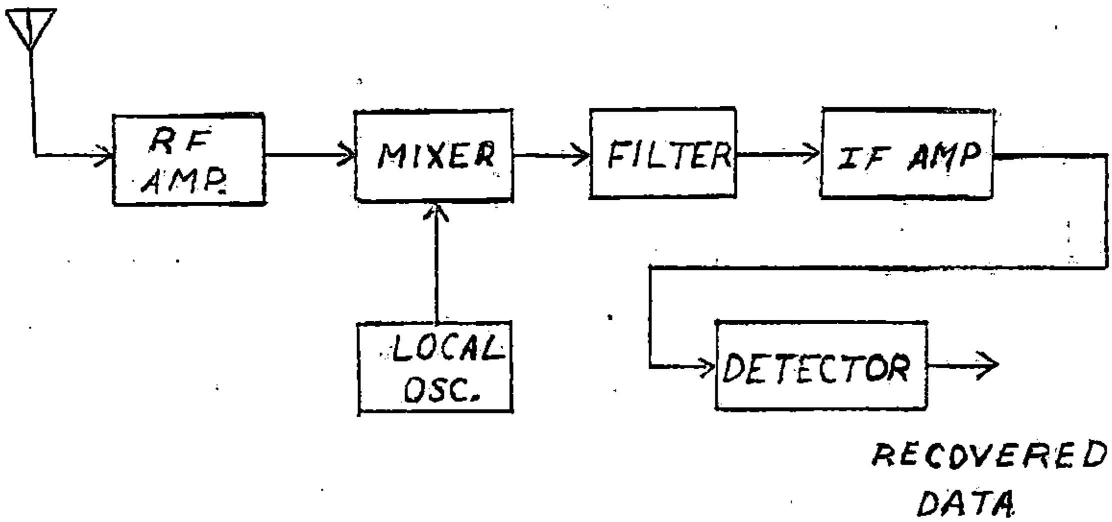


Figure 4. Block diagram of an AM radio receiver

is then amplified in the IF amplifier and finally the data waveform is recovered from this signal by some type of AM demodulator, such as an envelope detection circuit which determines the amplitude of the signal (that is, the data waveform).

The first circuit of the AM receiver to be experimented with was the input amplifier. The circuit which was used is a modification of the JFET VHF common source amplifier shown on page 291 of the Radio Amateur's Handbook (23) and is shown in Figure 5. (Modification of this existing circuit resulted in both a savings of time and a better amplifier than would have resulted had the amplifier been designed from scratch. This circuit modification approach was thus followed throughout the remainder of the research.) The construction and operation of this circuit are discussed in Appendix B. After adjustment of the input and output tuned circuits of the amplifier, a gain of approximately 16 db was attained at 150 MHz (150 MHz was chosen as the operating frequency because it is a clear frequency in Ames, Iowa.) This circuit will again be discussed in Section 4.

The second circuit of the AM receiver to be constructed was the local oscillator. Since a very frequency-stable receiver was desired, it was decided to build a common-base type crystal oscillator very similar to the one shown on pages 96 and 97 of the crystal oscillator book by Frerking (12). This oscillator was supposed to operate at 149.9 MHz so that the intermediate frequency would be low enough so that micropower op amps (see Appendix A) could be used as the heart of the IF amplifier. However, this oscillator was never made to operate at the crystal frequency during this part of the experiment, mainly due to a lack of understanding

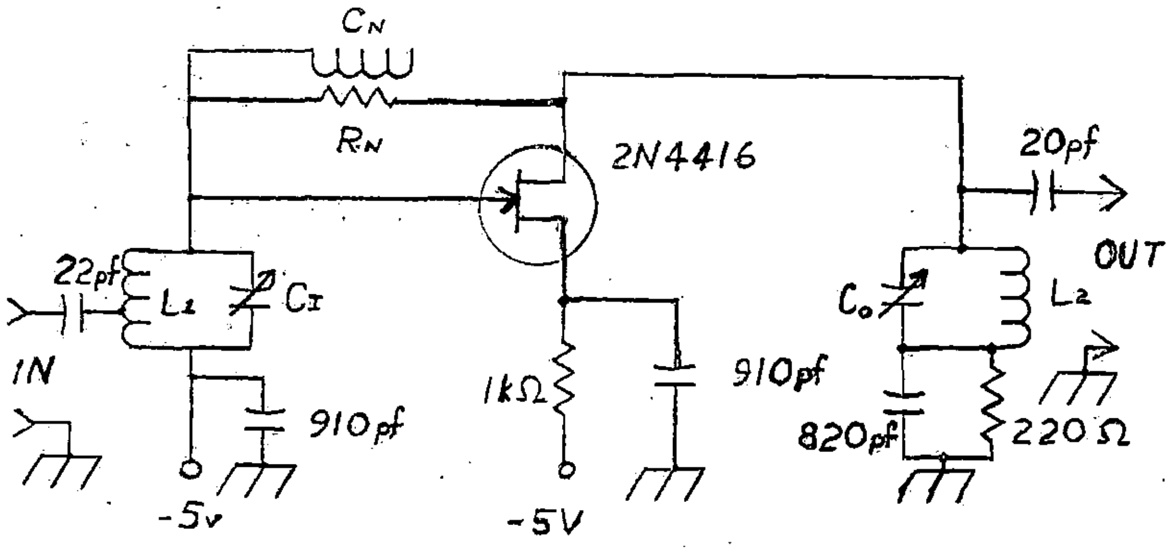


Figure 5. Common source JFET VHF RF amplifier (from page 291 of Reference 23)

of how the crystal and crystal oscillator circuitry function. Appendix C gives a brief description of how a crystal oscillator functions. For a more detailed description, see Frerking (12). For a description of the construction and tuning of a crystal oscillator, see Appendix D.

After a lack of results with the crystal oscillator, work was begun on the mixer stage of the AM receiver. However, shortly after this work was begun, a conference of the Solid State Affiliates was held at Iowa State University. After a presentation of this device at the conference, it was stated by a number of representatives at the conference (unfortunately, the names are not known) that the AM approach is probably not the best. They stated that low-power components for FM receivers (such as pager components, CMOS PLLS, etc.) are readily available and that FM communications have some major advantages over AM communications. For example, AM transmissions are more affected by noise than are FM transmissions. Also, AM signals tend to fade; thus, the true biological signal(s) being received may be distorted by the change of strength of the transmitted signal which may occur as the animal is moving toward or away from the central terminal unless precautions are taken to correct for this. FM signals, on the other hand, will either be detected or not detected by the receiver since the data is encoded in the frequency rather than in the amplitude of the carrier signal. The FM approach was thus determined to be the best. Again, modification of an existing device was the approach decided upon. At this point, the existing device which seemed to be the most appealing was a pager device, since the power consumption of this device is low and the size is small.

2. Pager Approach

The Motorola Pageboy II™ pager schematics and general information about this device were obtained from Mark Grootveld.[1] This pager draws only 3.8 mA from a 1.45 V battery when on standby (~5.5mW). The schematic of the circuitry of this device is shown in Figure 6. (A1, A2, A3 and A4 represent hybrid circuit modules.) This circuit is intended to function approximately as follows. The RF input amplifier (Block A1) receives the transmitted signal and amplifies it by 11.5 db. This signal is then mixed with a local oscillator signal (Block A2) of frequency such that the difference (transmitter carrier frequency - local oscillator frequency) signal has a frequency of 17.9 MHz. This signal is then amplified by about 23.5 db in module A2. At the output of the high conversion module (Block A2), the 17.9 MHz signal is passed through a six-pole crystal filter in order to reduce the bandwidth of signals around 17.9 MHz which are also passed; this both improves selectivity and greatly attenuates images that might be produced when the 17.9 MHz signal is mixed with the 17.935 MHz local oscillator signal of module A3 to produce a 35 kHz signal output. The signal suffers 3 db loss through the crystal filter, but is amplified by 14 db in the low conversion module (Block A3). Finally, the 35 kHz signal is amplified by 14 db and FM-demodulated (Block A4) to retrieve the audio information which was originally transmitted.

It was decided originally that two identical receivers should be constructed from the Motorola Pager parts. The complete pagers would have cost approximately $250 each. Therefore, it was decided to order two

[1] Mark Grootveld is an assistant engineer for the Ames Laboratories, Ames, Iowa.

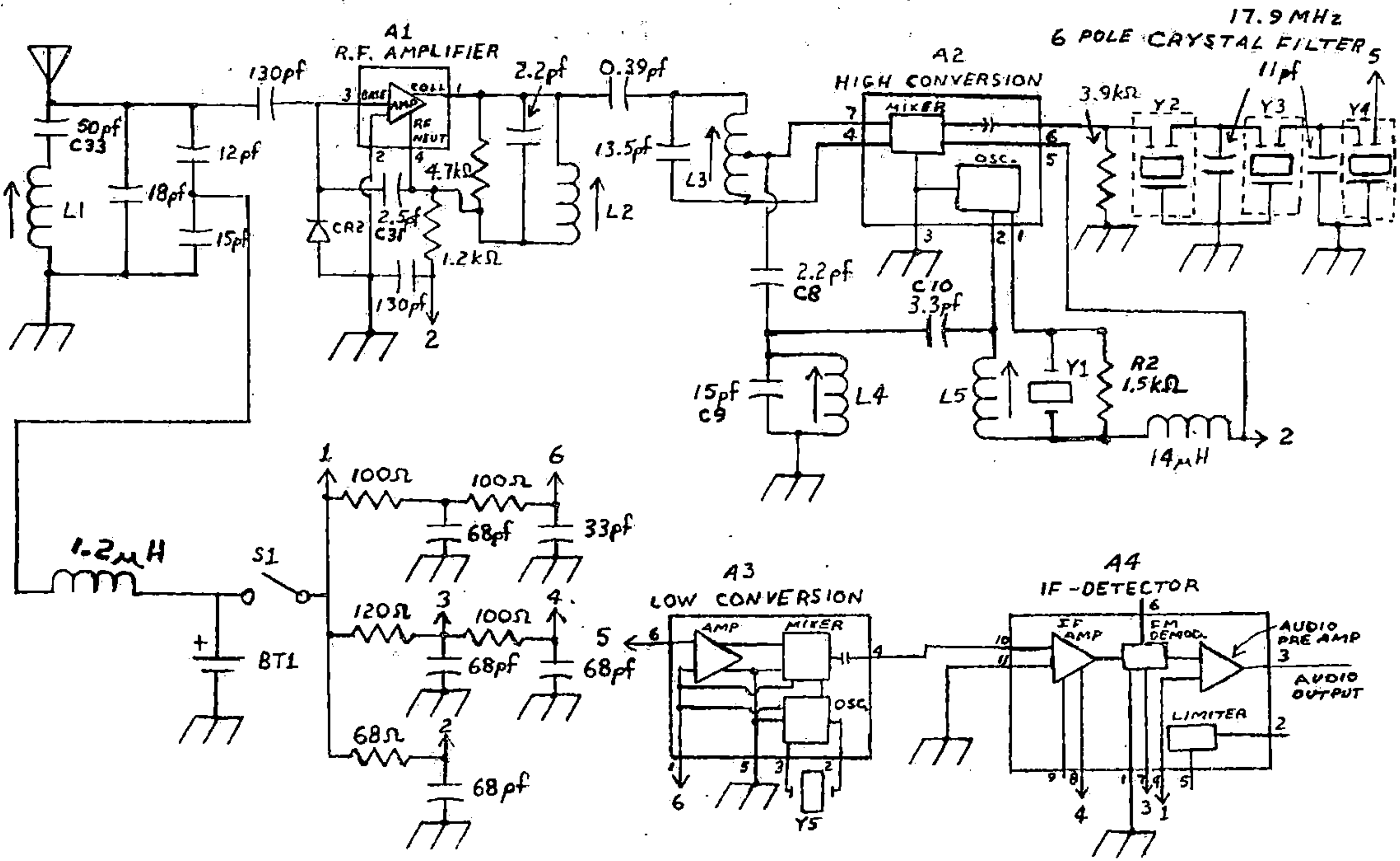


Figure 6. RF circuitry of the Motorola Pageboy II™ pager

hybrid modules of each of the modules A1, A2, A3 and A4 and two of the remainder of the RF circuitry shown in Figure 6. Unfortunately, it was not a simple task to make the receivers work once the parts were obtained. To begin with, when the parts were ordered, the six-pole filter between the high converter and low converter was not available. If this filter is not used, then there is no reason to use the low conversion module (A3). That is, without this filter, or one similar to it, an image frequency will amost certainly be present and the selectivity will be poor at the output of the low conversion module (A3). Thus, it would be just as well to use a higher local oscillator frequency in the high conversion module (A2) such that the signal output from this module has a frequency of 35 kHz. To improve the selectivity, this 35 kHz signal could be passed through an active bandpass filter. The approach to building an FM receiver from the pager parts was thus begun with the intent of using modules A1 and A2 of the pager circuitry and micropower op amps and CMOS PLLs in the IF amplification and FM detection circuitry. (See Appendix A for a description of these devices.)

The first part of this receiver which was constructed was the input RF amplifier. This was constructed by connecting discrete components to module A1 as shown in Figure 6. The maximum gain of this circuit was obtained at 145 MHz instead of at the desired 151 MHz. In order to try to move the maximum gain point to 151 MHz, C33 was reduced to 39 pF; however, this caused oscillations to occur. C31, the neutralization capacitor, was increased to 2.89 pF in an attempt to stop the oscillations. This stopped the oscillations and also caused the maximum gain point to increase to 146-147 MHz. It was discovered that by increasing C31 to 5.09 pF, the maximum gain point was increased to 151 MHz.

After the input amplifier stage was made to function, the next step was to construct the circuitry around the high conversion module (A2) and try to make the crystal oscillator in this stage oscillate at the crystal frequency. The oscillator in this module was designed to operate at one-third the frequency of the desired local oscillator signal. The manner in which the local oscillator signal is obtained is to tune the crystal oscillator so that it is overdriven, thus producing a distorted oscillator waveform rich in odd harmonics, and to filter out the third harmonic signal by using the tuned circuit, consisting of C9 and L4, tuned to resonate at three times the oscillator frequency. Assuming that the output signal of the high conversion stage is to be 35 kHz, and that the transmitted carrier frequency is 151 MHz, then the crystal oscillator should oscillate at $(151 + .035)/3 = 50.345$ MHz. The oscillator was never made to oscillate at the crystal frequency even by adjusting C8, C9, C10, L4, L5 and R2. Part of the problem could be that crystals not normally used in the device were being tried because an error was made in specifying the crystals on the original order. Since the parts took six weeks to arrive, it was decided that a new order for crystals would not be made and thus other crystals were tried.

It was hoped that by using an intermediate frequency of 35 kHz that the IF amplifier could be constructed from simple cascaded inverting amplifiers made from micropower op amps (such as the Fairchild μA776). (See Appendix A for a description of the μA776.) This would have reduced the cost of the receiver since each of the four modules in Figure 6 cost approximately $25.

After more frustration with the high conversion module (A2) and after consideration of the cost of the hybrid circuit modules, the pager idea was finally abandoned. The estimated cost of each receiver (using only modules A1 and A2 of the pager circuitry) would be a minimum of $60. For a research experiment which would incorporate a large number of these remote data acquisition devices (Figure 3), the receiver cost alone would be almost prohibitive. In addition, this $60 receiver would have rather poor selectivity due to the absence of adequate filtering.

After consultation with Joe Riley,[1] it was decided that the remote data acquisition device shown in Figure 3 would almost certainly have to remain external to the animal because of the large number of connections which would be made to the animal when monitoring a large number of biological signals. Since the device would be external to the animal, power consumption and physical size were not as confining restrictions as was originally thought. Therefore, it was and still is felt that there is a better approach to the receiver than using pager parts.

3. Superregenerative Approach

Joe Riley suggested the use of a modified commercially available FM receiver for the remote device. Before beginning this approach, it was felt that construction of a superregenerative device might be

[1]Joe Riley is an electronics engineer at the National Animal Disease Center, Ames, Iowa.

worth a try before going on to the modified FM receiver approach for the following reasons:

1. Information about superregenerative devices was found which indicated that the possible problem of receivers interfering, mentioned at the beginning of this chapter, might be overcome by incorporating a well-neutralized RF input amplifier prior to the superregenerative stage, to block superregenerative stage oscillations from reaching the antenna, and thorough shielding of this superregenerative stage, to prevent radio frequency radiations.
2. The proposed superregenerative receiver would be smaller than the modified FM radio device because it would require fewer components.
3. The proposed superregenerative receiver would consume approximately 1/2 the power of the modified Weatheradio™ of Section 4.

It was hoped that the superregenerative stage could be used as an FM demodulator. However, the manner in which this device operates makes it to be nearly an AM demodulator. The superregenerative stage functions approximately as follows. The signal from the antenna or from the input amplifier is fed into the superregenerator stage and amplified. The amplified signal is fed back into the input of the superregenerator stage and amplified again. The number of times that the input signal is amplified is determined by timing circuitry which controls DC bias to the transistor of the RF oscillator part of the stage. After the preset number of amplifications, the DC bias to this transistor is removed and then restored so that a new segment of the input signal can be amplified.

The timing circuitry which controls this DC bias can be adjusted so as to achieve the desired compromise between sensitivity and selectivity.

As was mentioned above, it was thought that by using a well-neutralized RF amplifier as an input stage to the superregenerator stage and adequate shielding of the superregenerative stage, that interaction between receivers could be prevented. The circuit which would have been used as the input amplifier is shown in Figure 5. (The construction and tuning of this amplifier are outlined in Appendix B.) However, it was not possible to neutralize this amplifier to the extent that would have been desired. This will be discussed more in Part B.

The next circuit to be constructed was the superregenerator stage. It was planned that FSK-AM transmission would be used such that a 400 Hz AM signal would represent a binary 0 and that a 1200 Hz AM signal would represent a binary 1. Thus, the output of the superregenerator stage would be either 400 Hz or 1200 Hz. (The audio amplifiers and demodulators which would have been used to convert the output signal of the superregenerator stage to a binary 0 or binary 1 are the same as were used for converting the output of the Weatheradio™ FM Detector in Section 4 and are shown in Figure 12.) In the actual design and construction of the superregenerator stage, the first approach was to build a slightly modified version of the common-base crystal oscillator shown on pages 96 and 97 of Frerking (12) with VCO output of the CMOS 4046 connected to the base of the oscillator transistor to provide the quenching signal (amplification control mentioned earlier); the quenching signal frequency could then be controlled by a voltage input to the VCO. (See Appendices C and D for construction details and function of the common-base crystal oscillator and Appendix A for construction details for using the 4046 as a VCO or PLL.)

The frequency range of the VCO is determined by two resistors and one capacitor connected externally to the 4046 and were initially chosen so that a frequency range of 30 kHz to 150 kHz could be used. The circuit of this superregenerator stage is shown in Figure 7. As can be seen, the input signal is inductively coupled into the tuned circuit of the common-base oscillator. The 910 pF capacitor, connected from point 1 to ground, acts as a low impedance to the transmitted carrier frequency but as a high impedance to the 400 Hz or 1200 Hz signals modulated on to the carrier signal. The tuned circuit (containing L2) acts as a high impedance to the transmitted carrier frequency and as a short to the low frequencies. The transmitted data can thus be recovered across the 910 pF capacitor. The oscillator was operated at 47.9 MHz, since crystals near this frequency are standard and available. The superregenerator stage was then tested with a transmitter also tuned to 47.9 MHz and amplitude modulated with a 400 Hz or 1200 Hz sine wave. The results were not successful, although the crystal oscillator functioned properly.

The second superregenerator device tried was a modification of the device on page 68 of the October 1969 issue of Radio Electronics (32). The unmodified circuit is shown in Figure 8a and the modified circuit is shown in Figure 8b. It was thought that the power consumption of the device in Figure 8a could be reduced by replacing the astable multivibrator quencher circuit with the CMOS 4046 VCO quencher circuit that was used in the device shown in Figure 7. However, using the same test

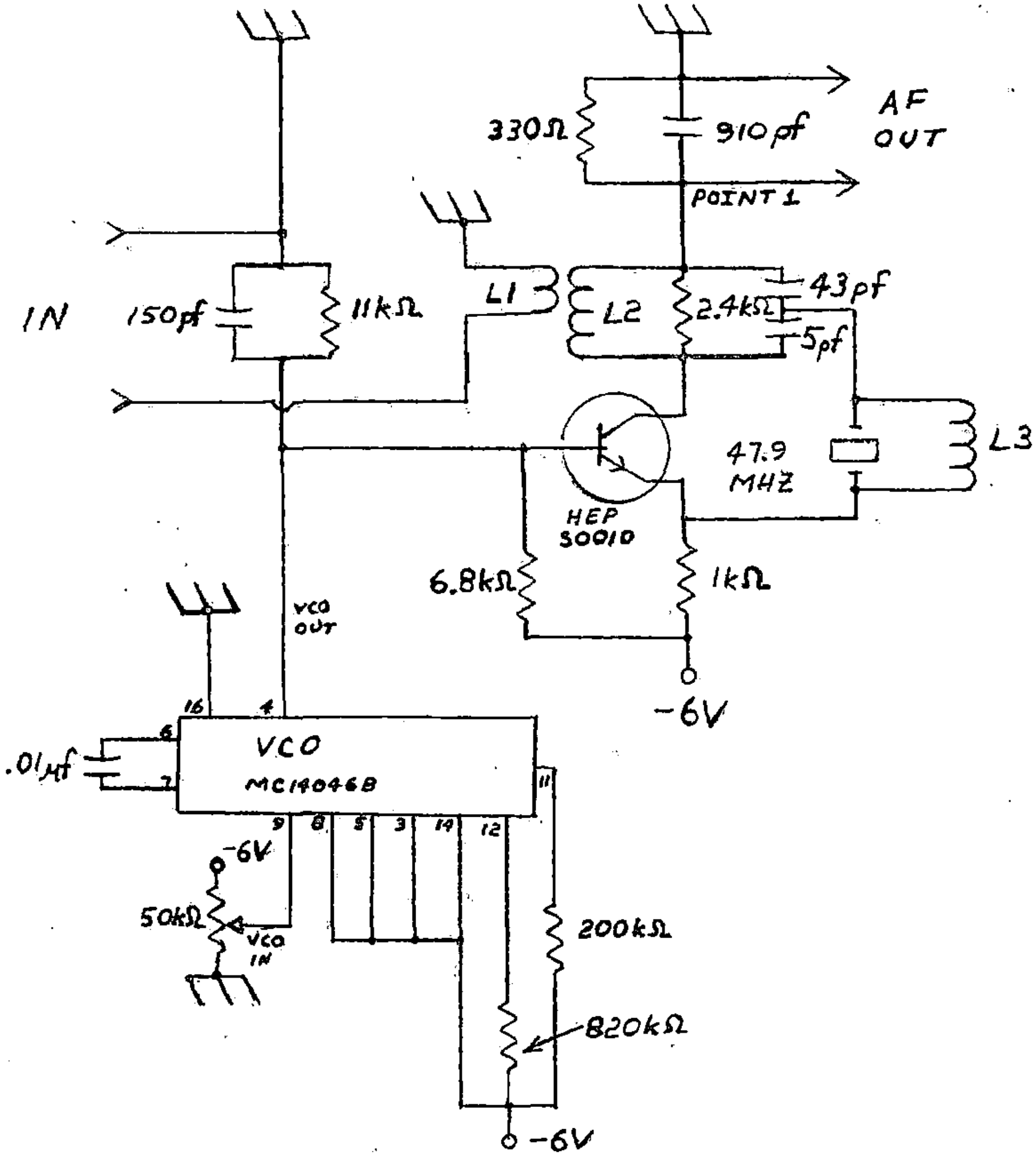


Figure 7. Superregenerative detector constructed from common-base crystal oscillator and CMOS 4046 phase locked loop

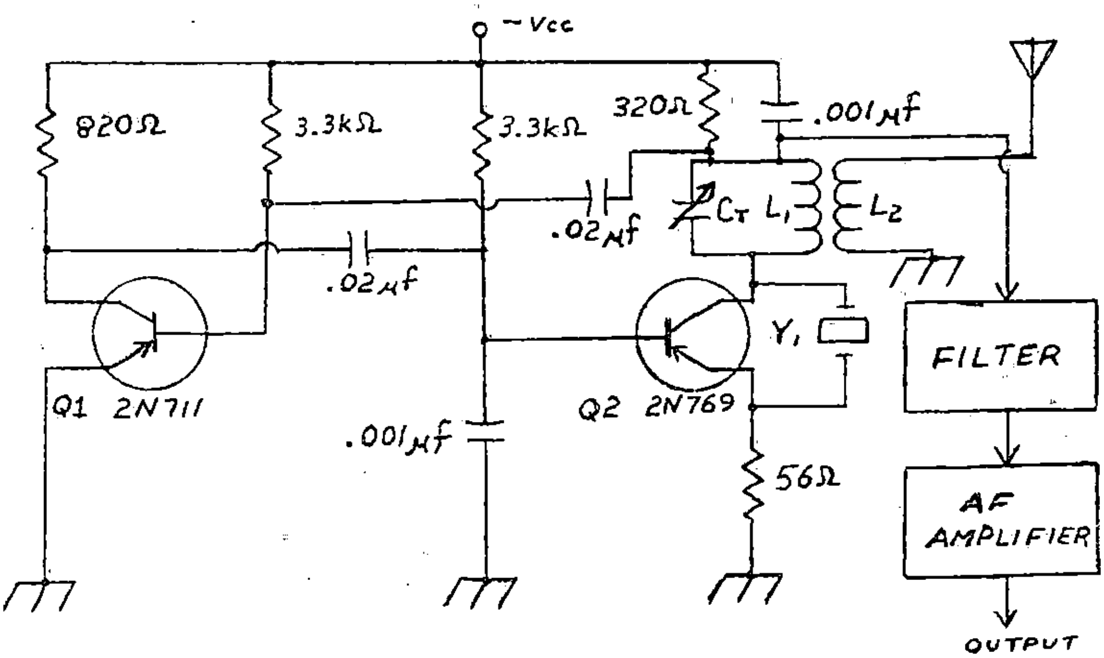


Figure 8a. VHF crystal controlled superregenerative detector from page 68 of Reference 32

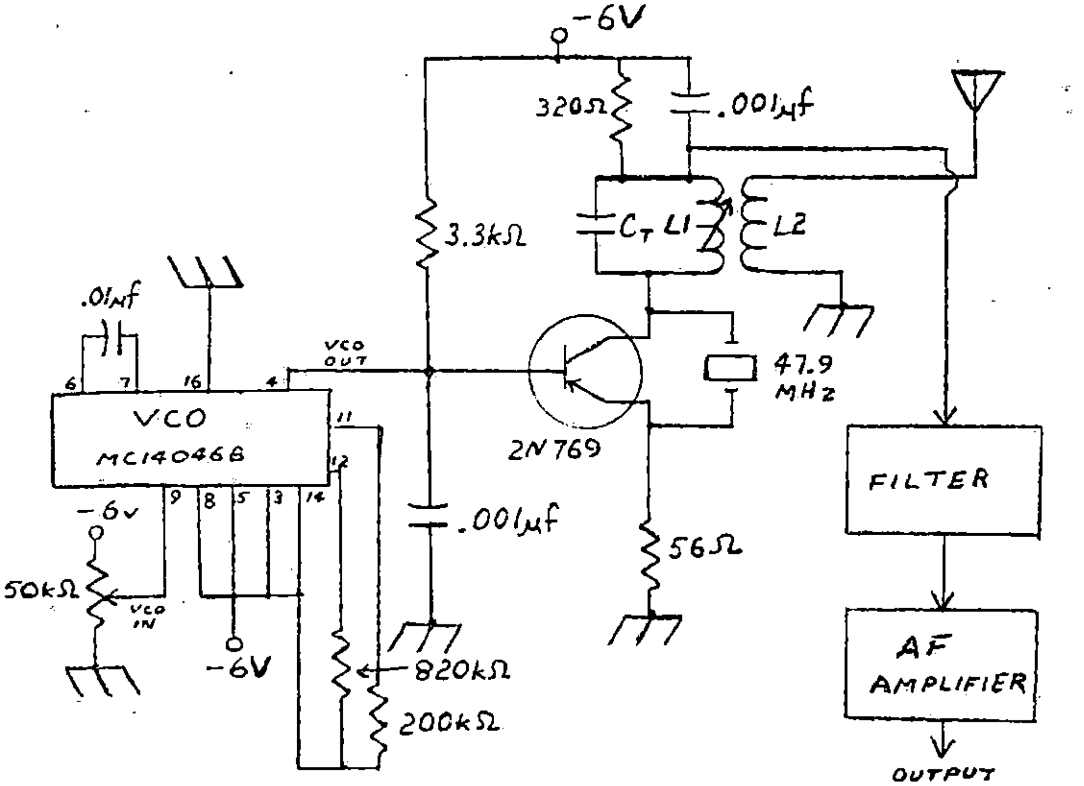


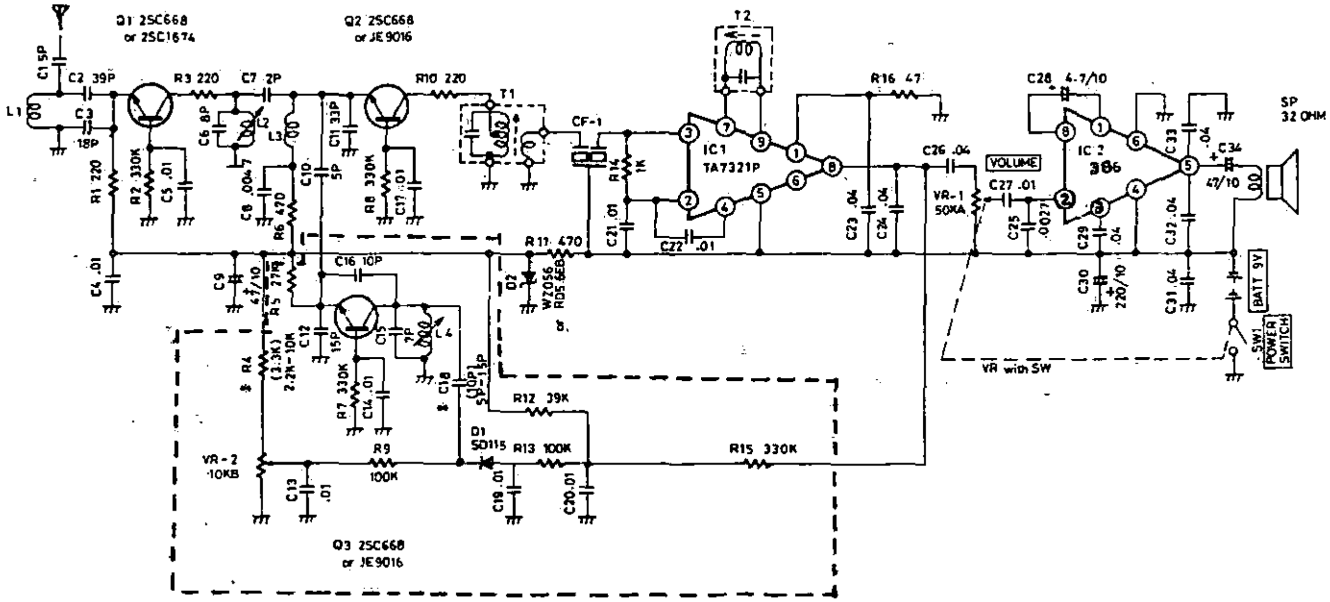
Figure 8b. Device of Figure 8a modified so that CMOS 4046 VCO replaces a stable multivibrator quencher circuit

transmitter as was used for the device shown in Figure 7, the superregenerator stage shown in Figure 8b did not function. With less than desirable results obtained with the input amplifier and the frustrations found in trying to make the superregenerator stage function, it was thought that time could be better spent in building a receiver from a modified FM radio. It still seems, however, that the superregenerative detector would show some promise as a receiver in a biological data acquisition system, due to its simplicity, if the problems encountered with the input amplifier and superregenerator stage could be overcome.

4. Weather Radio Approach

A search was made for an inexpensive FM receiver operating as nearly as possible to the desired 150 MHz. The device chosen is the Radio Shack Mini Weatheradio^{TM}. The schematic for this device is shown in Figure 9. Notice how simple this device is. It consists of an input RF amplifier stage, a variable frequency local oscillator, a mixer stage, a crystal filter at the output of the mixer stage, an IF amplifier/FM Detector IC, and an audio amplifier IC, the output of which feeds into the speaker. The device is designed to receive the frequency range of 162.4 to 162.55 MHz. (U.S. Weather Bureau stations broadcast continuous weather information on 162.55, 162.4 or 162.475 MHz.) The manner in which this device operates is approximately as follows. The received signal is amplified and then mixed with a local oscillator signal, the frequency of which is such that the output signal of the mixer is approximately 4 MHz. This signal then passes through a crystal filter, which greatly increases the selectivity of the device (reduces the bandwidth of signals around 4 MHz which pass

SCHEMATIC DIAGRAM



NOTE (1) ALL RESISTANCE VALUES ARE INDICATED IN "OHM" ($K=10^3$OHM, $M=10^6$OHM).
(2) ALL CAPACITANCE VALUES ARE INDICATED IN "μF" ($P=10^{-6}$μF).
※ MAY VARY FROM UNIT TO UNIT FOR BEST PERFORMANCE.

Figure 9. Weatheradio™ schematic

to the output of the crystal filter), to the IF Amplifier/FM Detector IC, the output of which is the audio information which was frequency-modulated onto the transmitted carrier. This audio signal is further amplified and fed into a speaker.

In order to make modification of the Weatheradio^{TM} as easy as possible, it was decided to operate the remote device near the weather band. In this way, the tuned circuits of the input amplifier would not need altering. The only alteration which would be necessary would be modification or replacement of the local oscillator. As a preliminary test of the Weatheradio^{TM}, the minimum operating voltage and the corresponding current drain were determined. At 5 volts, the radio was still fairly sensitive, but below five volts, the sensitivity dropped quickly, probably because the voltage became less than the minimum operating voltage for the IF Amplifier/FM Detector IC. Without the audio amplifier present in the circuit, it was found that the device draws only 8 mA at 5 volts ($\sim$ 40 mW).

The construction of the remote device receiver was begun by performing the only modification of the RF circuitry of the Weatheradio^{TM} necessary. This was done by removing the VFO circuitry (enclosed by dashed lines in Figure 9) and replacing it with a common-base crystal oscillator containing a frequency tripler output stage. (The operation and construction of common-base crystal oscillators is discussed in Appendices C and D.) The crystal used in the oscillator was constructed to be resonant at 53.51666 MHz. Thus, the local oscillator frequency used in the device is 3 x 53.51666 = 160.550 MHz, and the operating frequency of the receiver is approximately 160.6 + 4 = 164.6 MHz. The schematic of the Weatheradio^{TM}

circuitry with the VFO replaced by the crystal oscillator is shown in Figure 10. The idea for the frequency tripler stage used in this oscillator was taken from the crystal oscillator used in the high conversion module of the Motorola Pageboy II™ pager. This frequency tripler stage functions as follows. The tuned circuit of the grounded-base oscillator is tuned for crystal oscillation such that the output waveform becomes as distorted as possible. The third harmonic of the fundamental oscillation frequency is then present in the waveform and is removed by passing the signal through a bandpass filter tuned to three times the oscillator frequency and into the lower few turns (near ground) of the inductor of a parallel-tuned circuit tuned to three times the oscillator frequency. The output is then taken from the top of the parallel-tuned circuit through a high impedance (in this case, a 1 pF capacitor) and is nearly a pure third harmonic of the oscillator fundamental frequency. With a 50Ω load at the output of this tripler stage, a sine wave of $\sim$ 100 mV peak-to-peak amplitude at 160.550 MHz is output.

As was mentioned earlier, the operating frequency of this device is 164.6 MHz. Unfortunately, it was found that this device also receives the image frequency of 156.6 MHz. However, this has not caused any problems yet, and it does not appear that any such problems will result.

Before the remainder of the receiver circuitry was built, the LM386 audio amplifier IC was replaced as a convenience to the design process. The sensitivity of the circuitry shown in Figure 10 was then found to be $\sim$ 20 μV.

It was hoped that the next step of the remote device receiver construction would be to replace the IF Amplifier/FM Detector IC, which draws

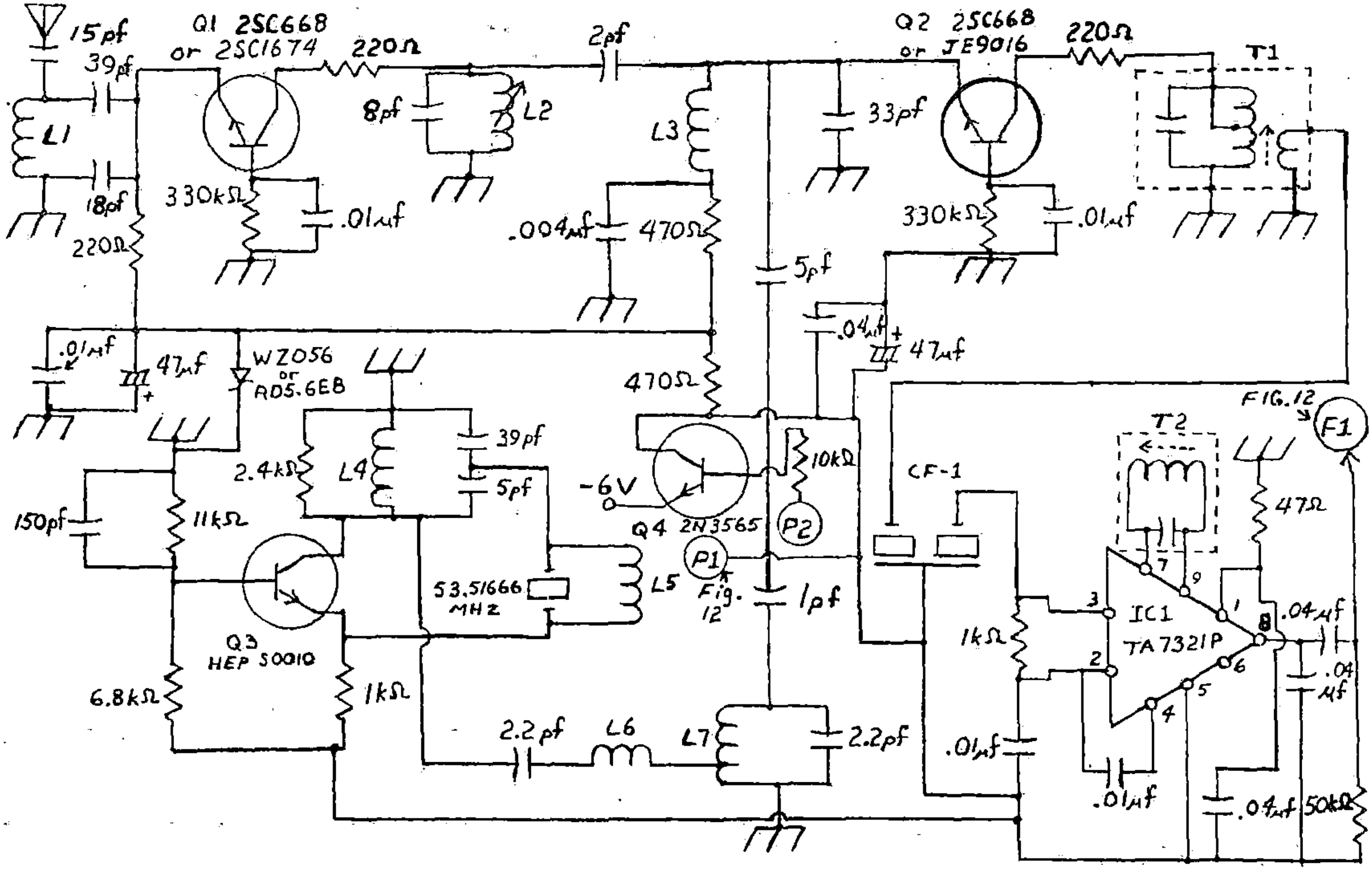
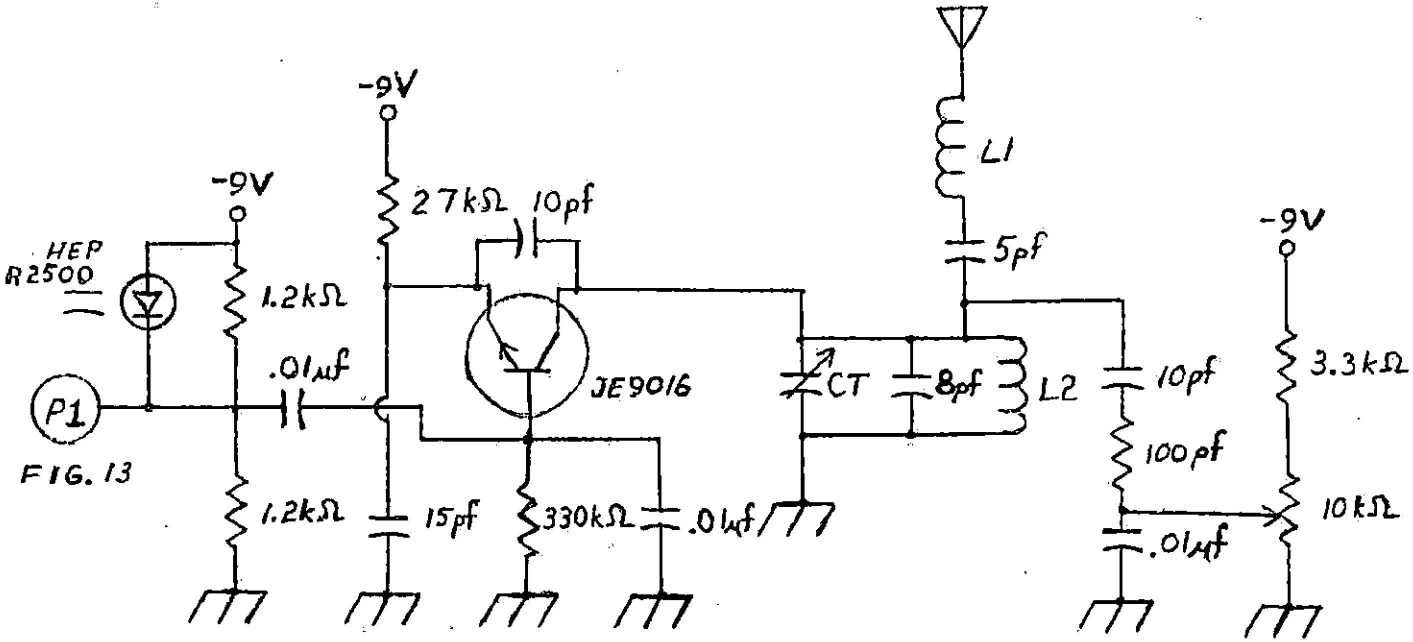


Figure 10. RF section of remote device receiver

4mA at 5 volts, with micropower op amps and CMOS devices. However, the micropower op amps and CMOS devices which are currently available will not operate at 4 MHz, the frequency at which the crystal bandpass filter also operates.

The next step in the design of the receiver was to rebuild the VFO circuit (which was originally the local oscillator in the Weatheradio™) as a test transmitter. This transmitter is shown in Figure 11. Notice at the base of the transistor of this VFO the presence of a voltage-variable capacitance diode which can be used to frequency-modulate the VFO. That is, when the voltage from the cathode of this diode to ground is varied, the frequency of the oscillator changes because the capacitance present across the diode changes. By applying a signal voltage from the cathode of the diode to the ground, the oscillator is frequency modulated. For this particular application, the linearity of the FM signal is not crucial. The thing that is crucial is that the frequency shift be great enough so that the FM Detector in the receiver circuitry can detect it. The frequency shift of the transmitter can be adjusted by varying the voltage amplitude of the signal applied across the voltage-variable capacitance diode. The test transmitter was first used with a simple sine wave input which was varied from about 400 Hz to 2 kHz. The 400 Hz to 2 kHz signal was clearly heard from the speaker of the receiver. This was an attempt to adjust the transmitter to the proper frequency and to determine if the frequency modulation technique was working. The next step was to FSK-FM modulate the transmitter since it was felt that FSK-FM transmission of binary data would be better than simple FM since it would be much more independent of transmitter carrier frequency



L1 - CONNECTED TO ANTENNA IN WEATHERADIO™ CIRCUIT

L2 - L4 OF FIG. 9 WITH CORE REMOVED

CT - 1/2" DIAMETER ALUM. DISKS APPROX. 1/4" APART

Figure 11. RF section of test transmitter

than would simple FM. The frequencies selected for the FSK were 400 Hz to represent a logic 0 and 1200 Hz to represent a logic 1. The FSK mode is also convenient in that the binary data can be stored directly on tape at the central terminal. The only complexity added to the receiver in using FSK-FM as opposed to simple FM is a CMOS 4046 PLL FSK demodulator IC. But, as will be mentioned in Part B, this same IC will also be used as the FSK modulator for the remote device transmitter. To FSK-FM modulate the test transmitter, the VCO of the CMOS 4046 PLL is used. The two resistors and one capacitor connected to the 4046 to control the range of frequencies from the VCO were selected such that a logic 0 input to the VCO would cause a 400 Hz signal output and a logic 1 input to the VCO would cause a 1200 Hz output signal. The output of the VCO was then connected from the cathode of the variable capacitance diode to the ground such that the test transmitter is FSK-FM modulated. To test this transmitter, the output of an astable multivibrator was connected to the input of the VCO. The alternating tones were heard at the receiver.

The next step of the design of the receiver was to construct an audio amplifier to amplify the FSK 400 and 1200 Hz signals from the IF Amplifier/FM Detector of the receiver. The LM386 IC which was used to drive the speaker was not satisfactory due to its high power consumption. (As was mentioned earlier, the LM386 was simply being used as a design aid.) The audio amplifier was constructed from a Fairchild μA776 micropower op amp (see Appendix A for description of this IC) as shown in Figure 12. The gain of this amplifier is $\sim$ 40 or 32 db below approximately 2000 Hz. Notice that the amplifier contains a low-pass filter with a frequency cutoff of $\sim$ 2000 Hz. This low-pass filter was used as an attempt to reduce the amplification of noise.

The output of this amplifier was then connected to the phase comparator input of a CMOS 4046 PLL (see Figure 12), and the two controlling resistors and capacitors of the 4046 VCO were adjusted so that the output of the PLL is logic level 0 when the phase comparator input is ≤ 400 Hz and logic level 1 when the phase comparator input is ≥ 1200 Hz. The output of this PLL could be passed through a Schmitt Trigger to the 87C48 microcomputer as a serial input. Any noise error detection which needs to be done could then be performed by the 87C48.

Again, using the FSK-FM modulated test transmitter mentioned earlier with an astable multivibrator at the input of the VCO, the output of the PLL at the receiver was following the transmitter astable multivibrator input.

At this point, the remote device receiver was complete. However, for purposes of testing and demonstration, it was decided that logic circuitry should be added to the receiver and to the test transmitter to determine the digital addressing capability of the receiver. The encoding circuitry at the test transmitter was made to be similar to what the actual central terminal transmitter address encoding circuitry would be. It was decided that the test transmitter should normally transmit a logic 0 (400 Hz-FM), and when data or address is to be sent, then a start bit (logic 1 or 1200 Hz-FM) should be transmitted just prior to the seven bits of data or address. Immediately after the seven bits of data or address are transmitted, the transmitter should again transmit a logic 0 continuously.

The encoding circuitry for controlling the data transmitter from the test transmitter is shown in Figure 13. Briefly, the circuitry

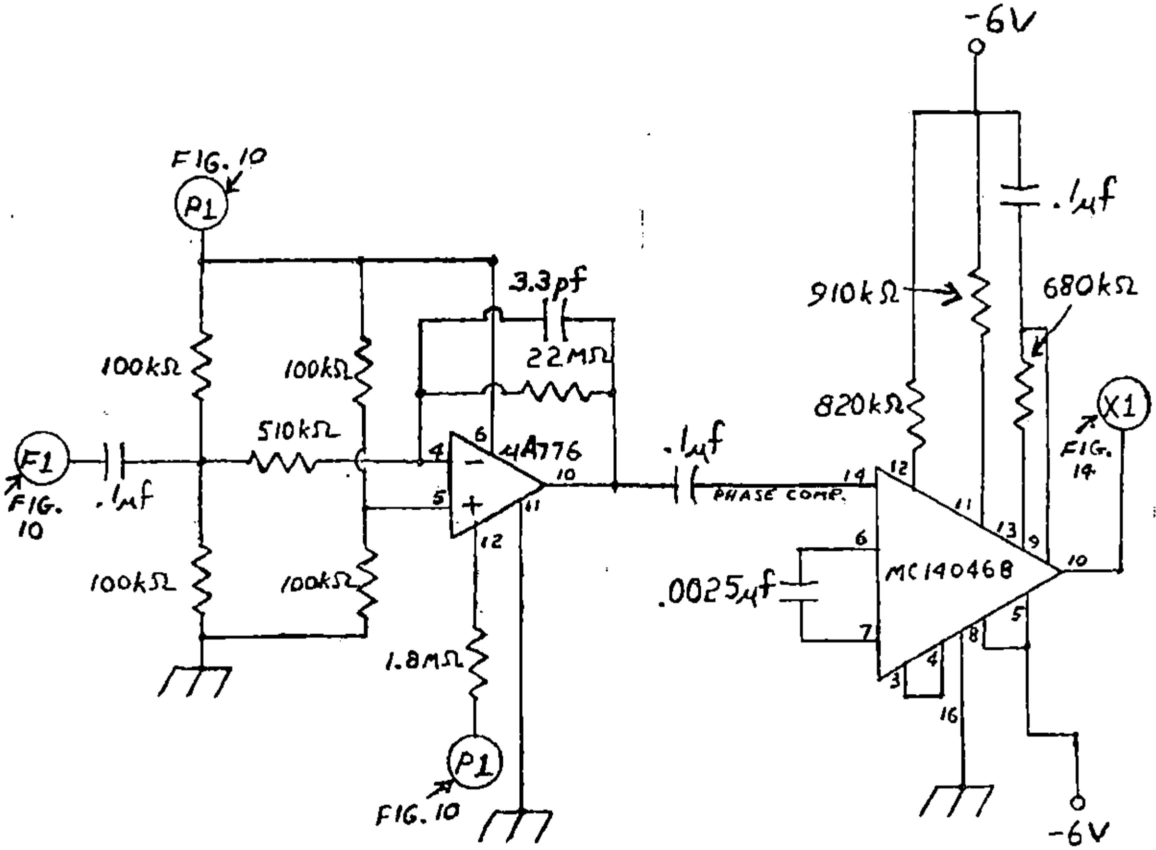


Figure 12. AF amplifier/FSK demodulator of remote device receiver

functions as follows. Before data can be transmitted, the reset switch must be depressed and released. When the data at pins 7, 6, 5, 4, 13, 14, 15 of the 14021 are to be transmitted from right to left (-9V represents logic 0 and ground represents logic 1), the transmit switch should be depressed and released. A logic 1 pulse then occurs at pin 9 of the 14021 which causes the data at pins 7, 6, 5, 4, 13, 14, 15 to be loaded into the 14021 shift register. Upon loading, a logic 1 is applied to the VCO input and represents the start bit. Each rising edge of the 12 Hz astable multivibrator thereafter, the data in the shift register shifts to the right, the previous rightmost bit being lost, and a logic 0 is shifted in at the left. (For a more detailed description of the circuitry shown in Figure 13, see Appendix E.)

The receiver decoding circuitry at the output of the FSK demodulation circuitry was, likewise, designed to mimic the manner in which the 87C48 microcomputer would interpret the data. The manner in which this decoding circuitry (shown in Figure 14) functions is approximately as follows. When a rising edge is detected at Point X1, a 10 ms logic 1 pulse is output by the monostable multivibrator. After this 10 ms pulse, if point X1 is still logic 1, then the 12 Hz astable multivibrator is enabled, and the transmitted data is shifted into the 14021 shift register. When the start bit reaches the rightmost bit of the shift register, the astable multivibrator is again disabled and the data bits output at pins 2 and 12 of the 14021 cause one of the four lights to come on, depending on the data bits transmitted. (Each light represents the identification code of a remote device.) The receiver reset momentary switch must be pressed and released after reception of

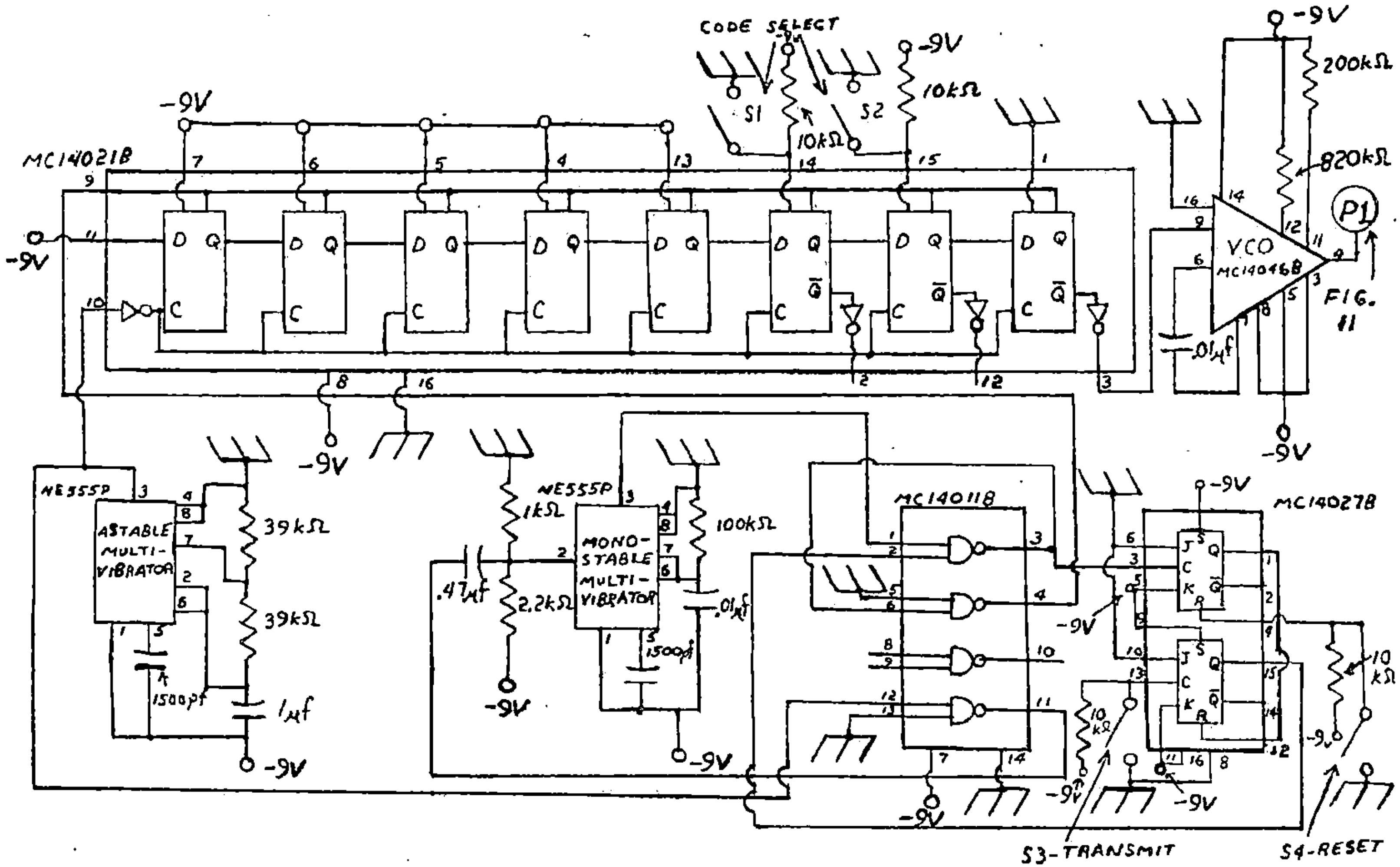


Figure 13. Address encoding circuitry of test transmitter

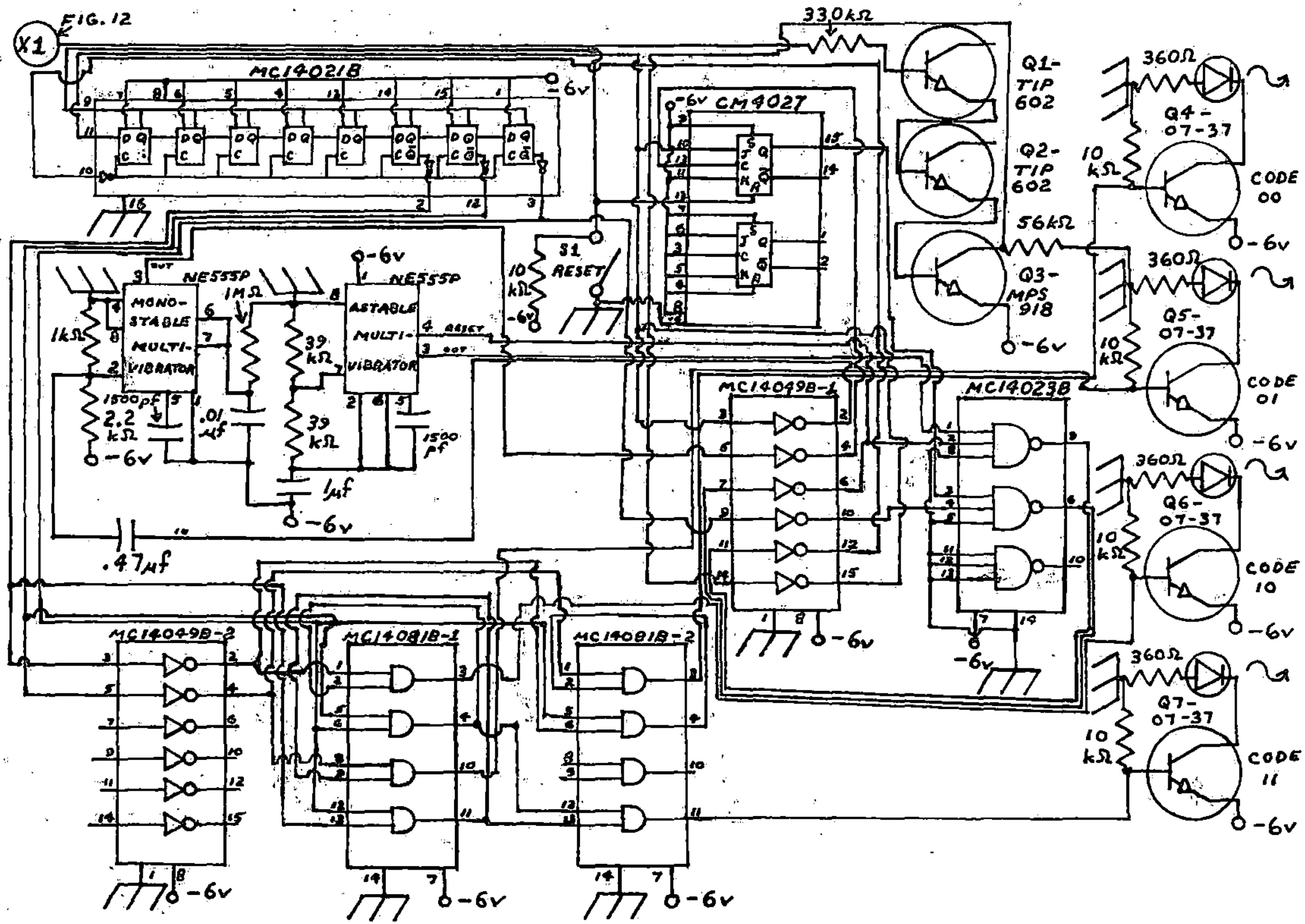


Figure 14. Address decoding circuitry of remote device receiver

any transmitted data before new data can be received. It was hoped originally that all seven data or address bits could be used as the receiver identification code, but five of the seven bits of the 14021 are not available. Thus, only two bits are used as the identification code. However, it was felt that this is adequate for experimentation and testing purposes. (For a more detailed description of the receiver decoding circuitry, see Appendix E.)

To test the test transmitter encoding and receiver decoding circuitry (at 12 Hz), each of the four possible identification codes was transmitted. If the received signal was strong, then the transmitted code was interpreted correctly at the receiver. If the received signal was weak, then noise caused errors in interpretation at the receiver. (The 87C48 microcomputer should be capable of detecting noise better than the circuitry shown in Figure 14.) However, a transmission rate of 6 Hz resulted in essentially no noise errors at the receiver.

B. Remote Device Transmitter Design

The final part of the research reported in this paper was to design the transmitter of the remote data acquisition device. This transmitter (shown in Figure 15) utilizes the VCO of the receiver FSK demodulator 4046 PLL (Appendix A discusses uses for the 4046) as the FSK modulator for the transmitter. The RF oscillator is a common-base crystal oscillator identical to the one used as the local oscillator in the receiver. Notice the voltage variable capacitance diode at the base of the transistor of the oscillator. When the voltage across this diode is varied, the oscillator frequency also varies because of the change of capacitance

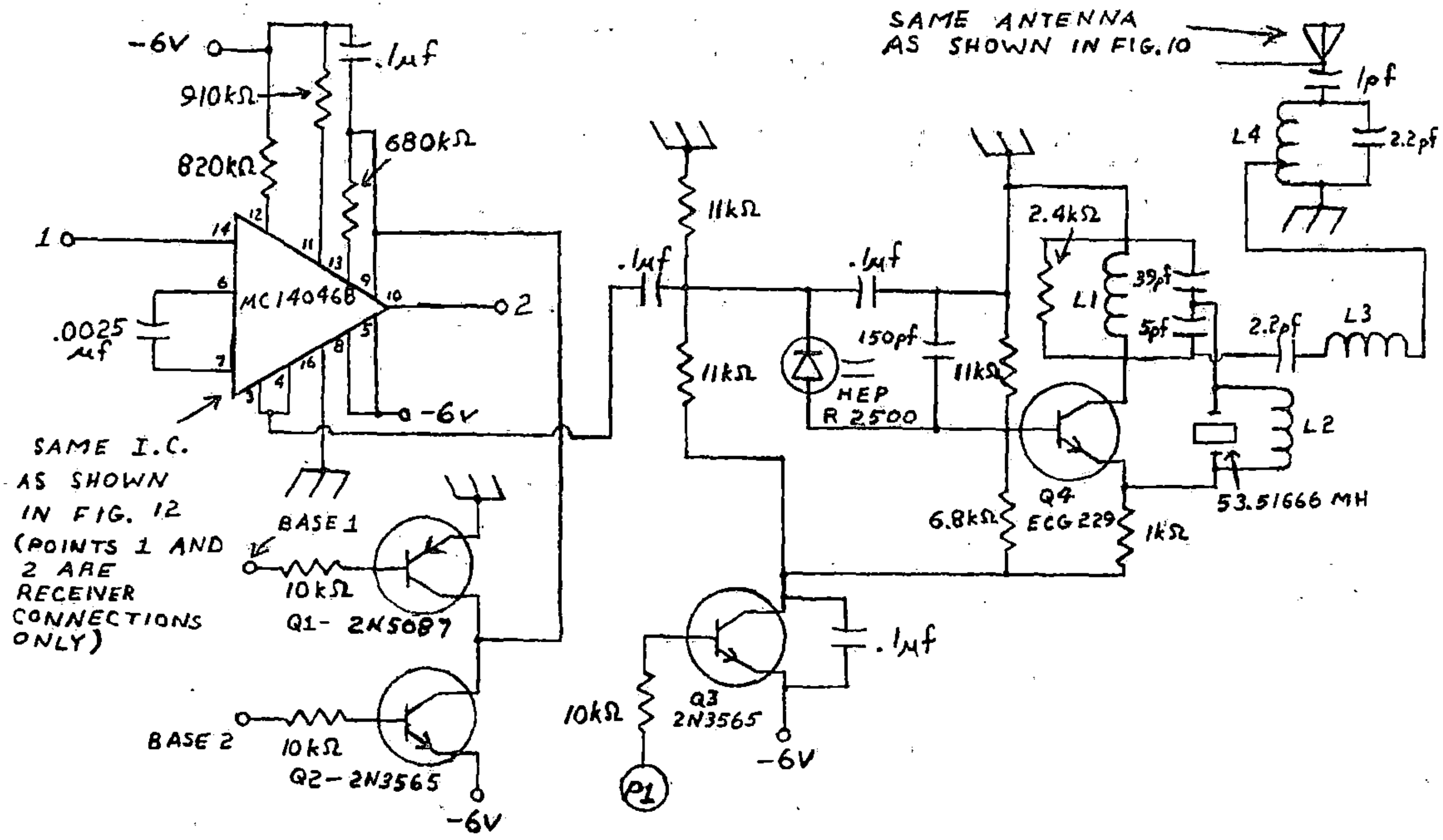


Figure 15. Remote device transmitter

in the oscillator feedback circuitry. Thus, when the FSK signal is applied at the cathode of this diode, the oscillator is FSK-FM modulated.

The transmitter FSK modulator 4046 (which is also the receiver FSK demodulator) requires two controls from the controller 87C48; one is required to control the output of the VCO so that it is 400 or 1200 Hz during transmission, and the other is required to switch the 4046 from the VCO mode for the transmitter to the PLL mode for the receiver or vice-versa. By connecting two transistors to pin 9 of the 4046 IC (as shown in Figure 15), and by proper control of these two transistors (by the 87C48 controller), the two controls, mentioned above, are achieved as follows. When Q1 is conducting (base 1 is at -6V) and Q2 is not conducting (base 2 is at -6V), the VCO input (pin 9) is at ground potential, causing the output of the VCO (pins 3,4) to be $\sim$ 1200 Hz. When Q1 is not conducting (base 1 is at ground potential) and Q2 is conducting (base 2 is at ground potential), the input to the VCO is at -6V, and thus the VCO output is a 400 Hz signal. Finally, note that when Q1 is not conducting (base 1 is at ground potential) and Q2 is not conducting (base 2 is at -6V), the 4046 will then operate as a PLL to the signal applied at pin 14. (More information about receive mode to transmit mode switching and vice-versa is given in Part C.)

This transmitter has not been constructed. However, no problems are anticipated in making it function since all of the components (4046 VCO, RF oscillator) were made to function earlier in the receiver design. From the results obtained from the local oscillator circuitry of the receiver, it is estimated that the transmitter circuitry will draw $\sim$ 2 mA at 6V and output $\sim$ 1 mW into the antenna.

It was hoped initially that the local oscillator of the receiver could also be used as the RF oscillator of the transmitter. However, a problem, which was not solved, was encountered which involved switching the antenna from the transmitter to the receiver or vice-versa for the appropriate mode. An attempt at electronic switching (shown in block diagram form in Figure 16) by use of a neutralized VHF JFET amplifier was made. The idea was that this switching amplifier should provide 0 db gain from input to output when the -6V power supply lead is connected to the amplifier (by means of a switching transistor) and 0 gain from input to output when the -6V power supply lead is disconnected. Thus, during the transmit mode, the -6V power supply lead would be disconnected from the RF input amplifier, the IF amplifier/FM detector IC, and from the μA776 micropower op amps used in the FSK demodulator circuitry. The FSK demodulator 4046 would be used to FSK modulate the local oscillator. The local oscillator signal would pass through the mixer, through the switching amplifier (with 0db gain) and finally to the antenna. During the receive mode, the -6V power supply lead would be disconnected only from the switching amplifier. It was hoped that the output of the local oscillator would be isolated from the antenna when the switching amplifier was "off." However, even when the switching amplifier was disconnected from the -6V power supply lead, a 100 mV input resulted in a 2 mV output. This 2 mV signal to the antenna would block even fairly strong signals from being detected by the receiver.

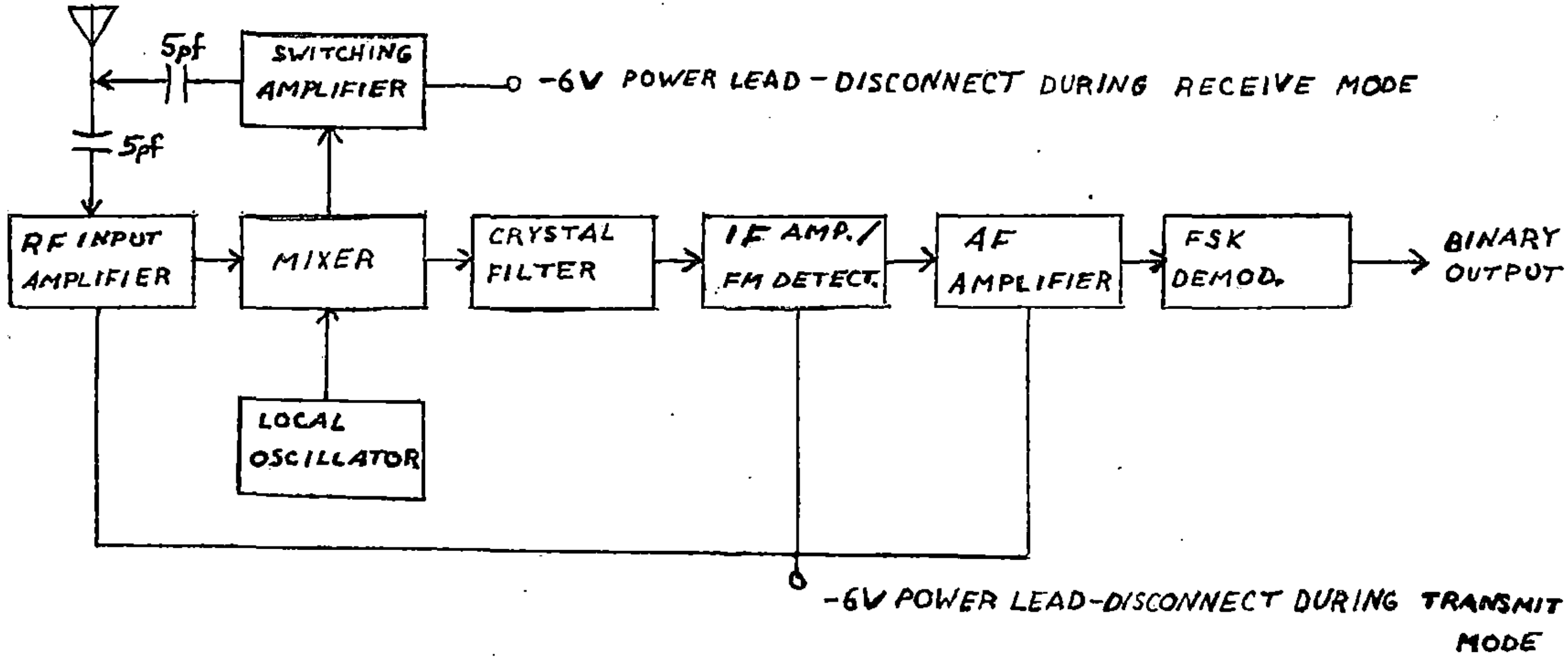


Figure 16. Block diagram of proposed remote device receiver/transmitter with antenna switching amplifier

C. Receive-to-Transmit Switching

Under standby operating conditions, the remote device receiver would normally be "on" and the transmitter "off". The transmitter RF oscillator is held "off" by the 87C48 by applying -6V at point P1 in Figure 15. The receiver circuitry is held on by the 87C48 by applying a ground potential at point P2 in Figure 10; this allows the -6V power supply lead to be connected to all parts of the receiver circuitry. When it is necessary for the remote device to go to the transmit mode, the 87C48 applies -6V to point P2 in Figure 10 in order to remove power from all stages of the receiver, except for the FSK demodulator 4046 PLL, and applies a ground potential at point P1 in Figure 15 to apply power to the transmitter RF oscillator. Data transmission can then occur upon proper control of the 4046 VCO discussed in Part B.

VI. CONCLUSIONS

The conclusions discussed here do not include comparisons between current telemetry devices and the proposed remote data acquisition device. The conclusions for these comparisons were fairly well-stated in Chapters III and IV. Rather, the conclusion concentrates on the experimentation which was done with the receiver and transmitter portions of the remote data acquisition device.

The modified Weatheradio™ appears to have the best sensitivity and selectivity, for the price, of any of the four receiver devices discussed in Chapter V. This receiver has a sensitivity of $\sim$ 20 μV and costs less than $20. A price is paid in the power consumption. This device draws approximately 8 mA at 6V (48 mW), which would allow four D-size batteries to operate the receiver for nearly two days. Assuming that the remainder of the remote device circuitry will consist of very low-power devices and that the remote device will be externally mounted on the animal, a set of four Nicad D-size batteries could be used to power the device and exchanged every day or two for freshly charged batteries. Even with the inconvenience of having to change batteries often, this receiver device is still more desirable than the other three considered and would almost certainly be satisfactory as the remote device receiver.

The AM receiver approach is not considered to be a promising remote device receiver due to the noise and signal fading characteristics of amplitude modulation.

Likewise, a receiver consisting of pager parts is not considered to be a promising remote device receiver due to the high cost of the pager parts. Even for a poorly selective receiver utilizing only modules A1 and A2 of the RF pager circuitry shown in Figure 6, the cost of each receiver would be ≥ \$60. If all of the RF pager parts were used, the receiver would cost ≥ \$100. The only advantage of the pager approach is its very low power consumption.

The superregenerative approach, however, still shows some promise as a remote device receiver due to its low power consumption and simplicity, but the difficulties found in trying to construct a well-neutralized amplifier and making the superregenerator stage operate must be overcome.

Finally, a low-power transmitter consisting of a 4046 VCO and grounded-base crystal oscillator seems to be satisfactory as an FSK-FM remote device transmitter. This device consumes approximately 2 mA at 6V (12 mW) and outputs approximately 1 mW to the antenna.

Thus, by using the modified Weatheradio™ as a receiver and the device mentioned above as a transmitter for the remote data acquisition device, the groundwork is laid for further design work on the remote data acquisition device.

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VIII. ACKNOWLEDGEMENTS

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IX. APPENDIX A

A. Low-Power Integrated and Monolithic Circuits Useful in Biotelemetry

The Motorola CMOS Data Book (20) presents many CMOS logic ICs and some CMOS analog ICs, which consume less than 100 μw apiece, which might be useful in biotelemetry. However, no attempt is made here to examine all of these devices. The usefulness of some of these CMOS logic ICs is apparent from the data encoding circuitry of the test transmitter (Figure 13) and from the data decoding circuitry of the modified Weatheradio™ receiver (Figure 14). Some information about these CMOS logic ICs is presented in Appendix E. The only CMOS device discussed in this appendix is the CMOS 4046 phase-locked loop. Another low-power device which will be discussed is the Fairchild μA776 micropower op amp (see Reference 18). Finally, a brief discussion will be given on a very low-power command receiver which can be used to turn power on and off to remote circuitry.

1. The MC14046B as a Phase-Locked Loop

The block diagram of this IC (from page 7-124 of the Motorola CMOS Data Book (20)) is shown in Figure A1. Pin 16 (VDD) is the positive power supply connection, and Pin 8 (VSS) is the negative power supply connection; the potential which can be applied between pins 8 and 16 can vary from 3 to 18 volts. (This is fairly typical of all series 4000 CMOS logic devices.) Pin 5 is an inhibit input, but for the uses in this research, the 4046 was always enabled, and thus pin 5 was always placed at potential VSS.

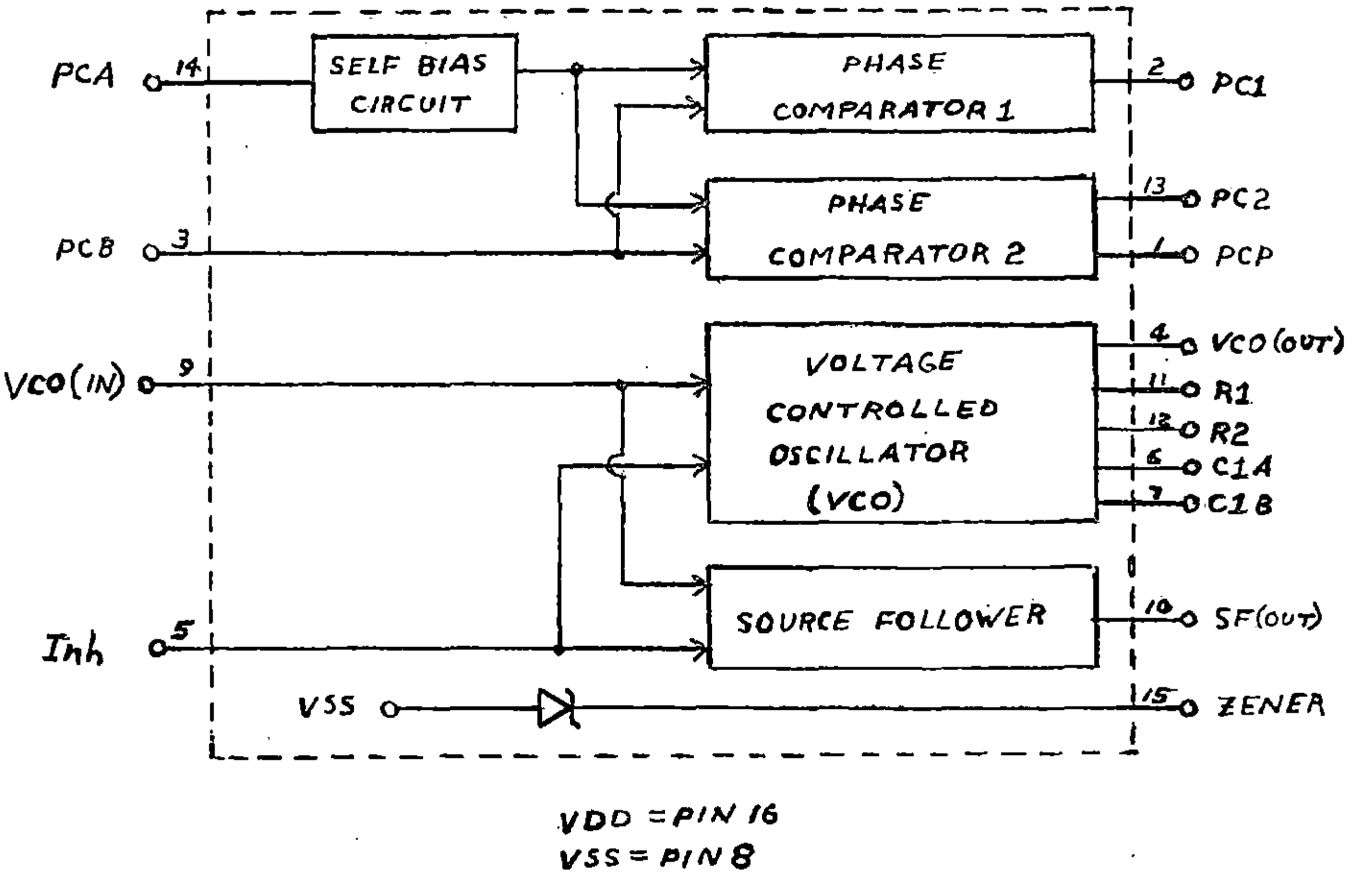
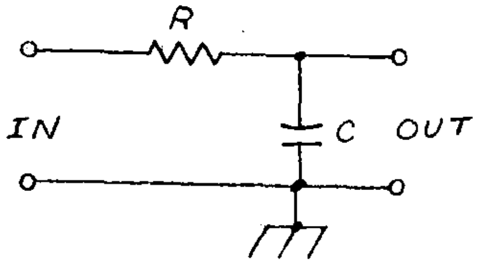


Figure A1. Block diagram of MC14046B phase-locked loop IC (from page 7-124 of Reference 20)

When this IC is connected as a phase-locked loop as shown in Figure A2, the input signal is applied to PCA (pin 14). When the frequency of the input signal lies between f(min) and f(max), (f(min) $\approx$ 1/(R2(C1+32pF)) Hz and .7 MHz > f(max) $\approx$ 1/(R1(C1+32pF))+f(min) Hz where 10k$\Omega \leq$ R1 $\leq$ 1MΩ , 10k$\Omega \leq$ R2 $\leq$ 1MΩ , and 100pF $\leq$ C1 $\leq$ .01μF), the voltage output at pin 10 (pin 10 is the output of a source follower, the input of which is connected to pin 9) is nearly linear with respect to the frequency of the signal input at PCA. For a more detailed discussion on how the PLL functions, see Reference 20.

The low-pass filter shown in Figure A2 should be constructed as described on page 7-128 of the Motorola CMOS Data Manual (20). One filter described is a simple RC network as shown here.



R and C should be chosen so that

RC = 1/(2πf(C)) where f(C) = (f(max) - f(min))/2 .

2. The MC14046B as a Voltage Controlled Oscillator

Notice that by disconnecting pin 9 (Figure A2) from the external low-pass filter and pin 4 from pin 3 that the VCO of this IC can be used apart from the remainder of the device. As the voltage input at pin 9 is varied from VSS to VDD, the frequency of the signal at pin 4 will vary from frequency f(1) to f(2), f(1) and f(2) being dependent on the selection

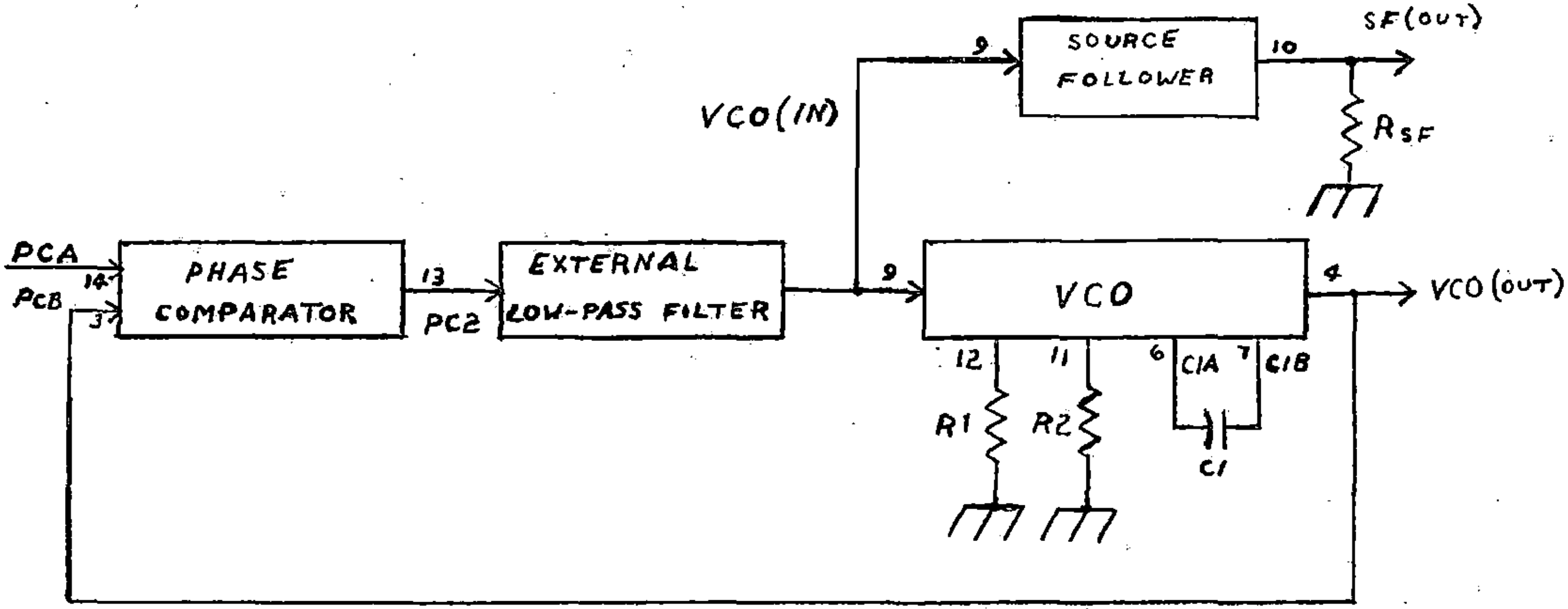


Figure A2. Block diagram of MC14046B being used as a phase-locked loop (from page 7-128 of Reference 20) of R1, R2 and C1. Approximate values for these components can be determined from the equations given in Section 1A for f(min) and f(max).

3. The Fairchild μA776 Micropower Op Amp

In connecting the μA776 micropower op amp IC into an amplifier circuit, it is connected in nearly the same way as any other op amp. The only pin of this IC which is not found on most other op amps is the set current pin. A resistor is connected between this pin and the negative power supply lead (VCC(-)) to adjust the quiescent current drawn by the device. On page 12-127 of the Fairchild Linear Integrated Circuits Data Book (18), it is stated that the set resistor (R(SET)) should be selected according to the equation I(SET) = (VCC(+) - .7 - VCC(-))/R(SET). I(SET) is usually varied between 1 and 100 μA and controls the gain-bandwidth product. The frequency used for this device in this research is so low that I(SET) is not critical, and thus it was chosen to be very small (R(SET) is 1.8MΩ, so I(SET) = 3 μA).

The offset null pins of the IC need not be used. The noninverting and inverting inputs and the output of the IC are connected to produce an inverting amplifier as would be constructed from any other standard op amp. (This op amp connected as an inverting amplifier is shown in Figure A3.) The positive and negative power supply connections were also made in the same manner as for a standard op amp. The supply voltage can be varied from ± 1.2 V to ± 18 V. (In this research, ± 3 V was used as the power supply voltage.)

4. A Monolithic Micropower Command Receiver

A device which was not used in this research but which is quite attractive as a possible power-saving device is a monolithic micropower command receiver (described on page 42 of the February 15, 1971, issue of Electronics (10)) which can extend the lifetime of an implanted telemetry device to almost the shelf life of the battery. When the proper radio frequency signal is transmitted to this device, it turns power on or off to the remainder of the telemetry circuitry. This device was developed by P. H. Hudson and has a sensitivity of > 100 mV and power dissipation of < 15 μW from a 1.35 V mercury cell.

One other device, which was described on pages 124, 125 of the March, 1972, issue of Wireless World (33), switches the receiver supply rail off and on with a 10% duty cycle signal from the squelch circuit. When a signal is received, full power is applied to the receiver supply rail.

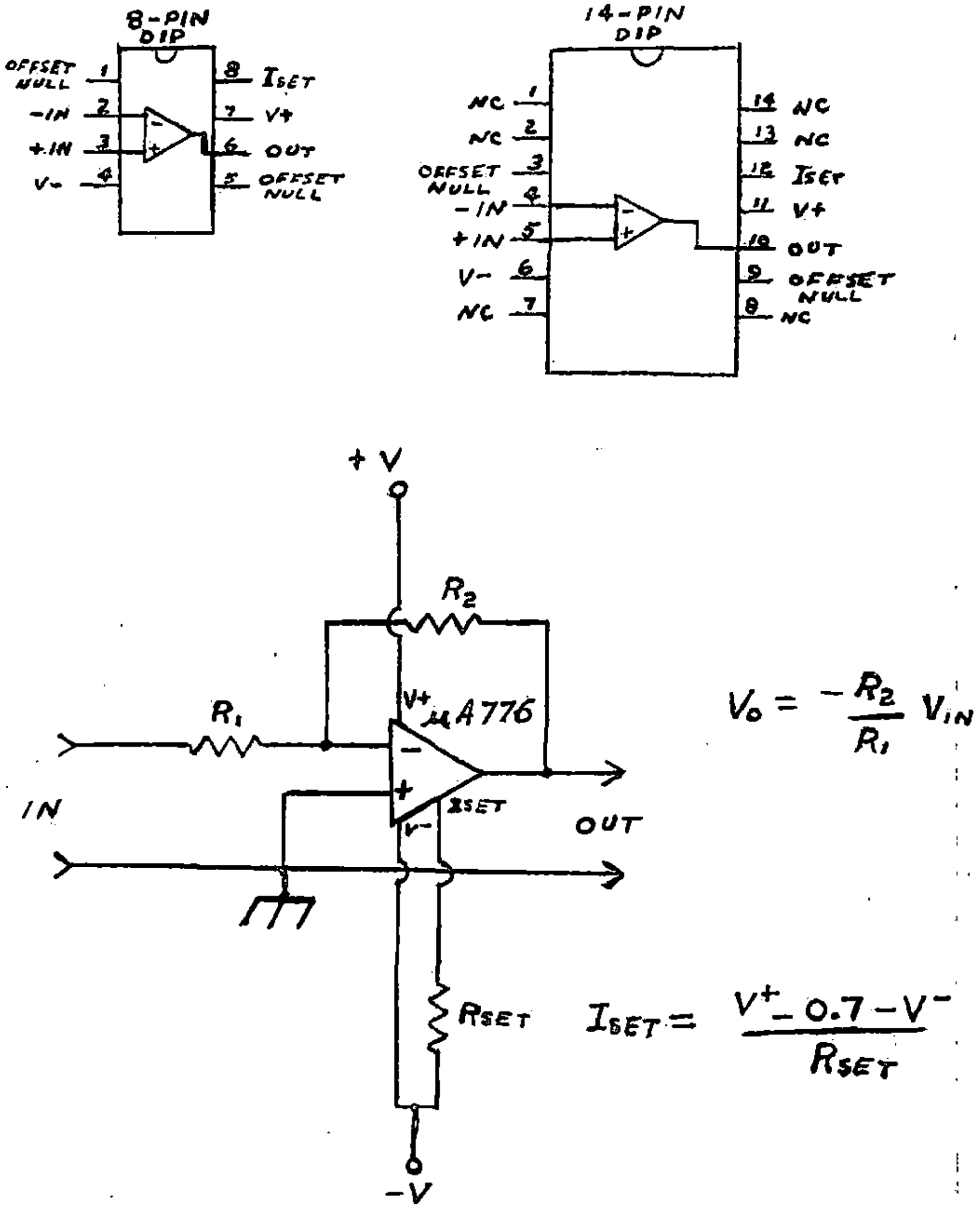


Figure A3. Pinout for Fairchild µA776 micropower op amp IC (from page 12-120 of Reference 18) and connection of the µA776 as an inverting amplifier

X. APPENDIX B

A. Construction and Tuning of the JFET VHF Amplifier Shown in Figure 5

The JFET VHF amplifier shown in Figure 5 and again in Figure B1b is a modification of the JFET VHF amplifier shown in Figure 9-1B on page 291 of the 1976 Radio Amateur's Handbook (23) (Figure B1a). The construction details for making the inductors (L1 and L2) of the input and output tank circuits of the amplifier were taken from the description of the input RF amplifier circuit shown in Figure 32 on page 9-20 of the 1980 Radio Amateur's Handbook (24).

1. Construction Steps

The amplifier shown in Figure B1b was constructed according to the schematic shown in Figure B1a with the following exceptions:

1. The source resistor was increased from 270 Ω to 1kΩ so that the quiescent gate-to-source voltage would be more negative and, thus, the quiescent drain current would be smaller. Also, the drain resistor was decreased from 470 Ω to 220 Ω so that the quiescent drain-to-source voltage would be increased.
2. A positive ground was used instead of a negative ground so that the circuit would be consistent with the other circuits with which it is used.
3. The 100 pF capacitor and the 47kΩ resistor of the gate circuitry were omitted. With the 100 pF capacitor in the gate circuit, the 47kΩ resistor is necessary so that the gate is held to the

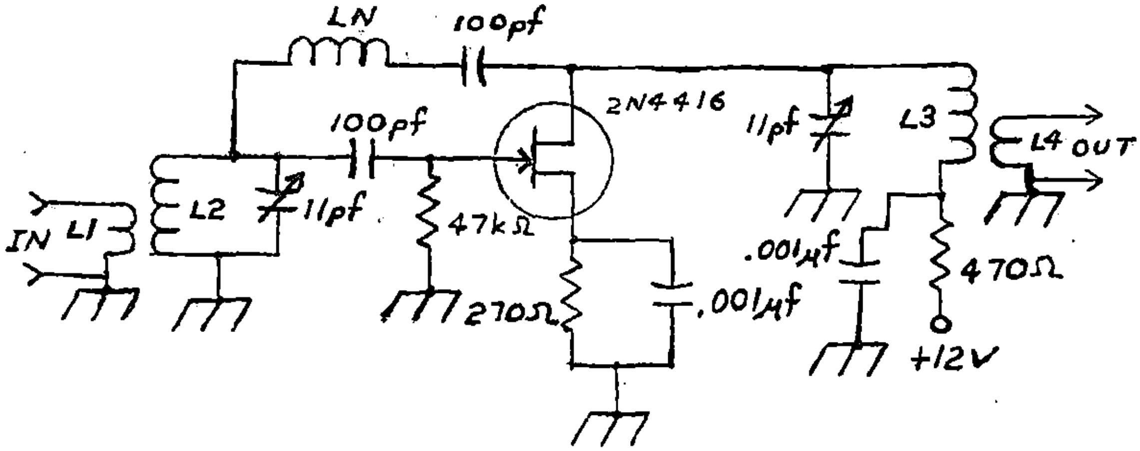


Figure B1a. 145 MHz common-source JFET VHF amplifier from page 291 of Reference 23

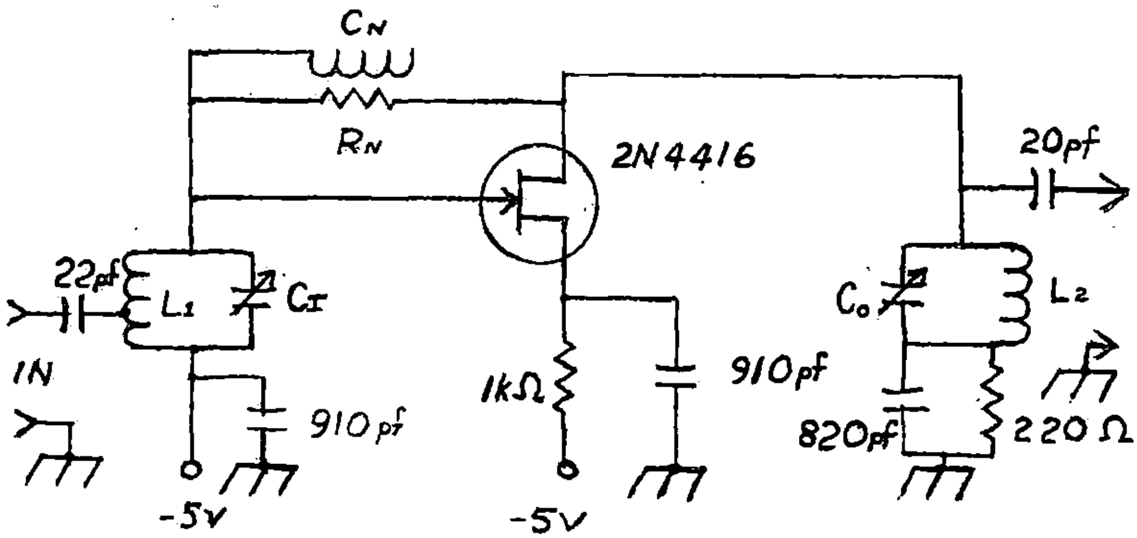


Figure B1b. 163.5 MHz common-source JFET VHF amplifier resulting from modification of the circuit in Figure B1a negative power supply potential. The reason for the 100 pF capacitor is not certain.

4. Capacitive coupling to the input and output tank circuits was used rather than inductive coupling, since inductive coupling is more dependent on component placement than is capacitive coupling.

5. The neutralization components were selected so as to allow maximum gain for the amplifier and consist of R(N) and C(N) rather than L(N). (The selection of these components is discussed more in the following discussion about tuning.)

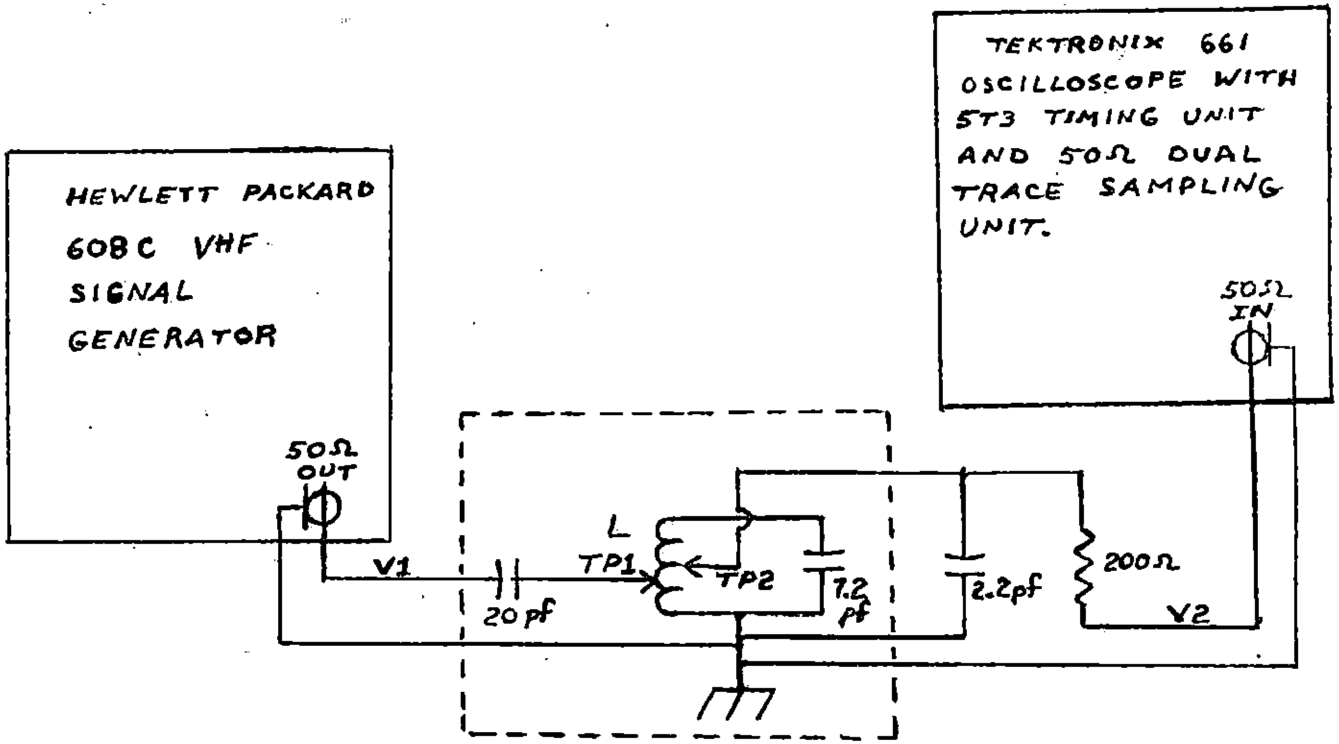
Both L1 and L2 were constructed by wrapping five turns of No. 20 enamel wire on a 1/4-20 bolt using the threads as a guide. This same inductor was used with a 10 pF capacitor to form the input tank of the 144 MHz input RF amplifier shown in Figure 32 on page 9-20 of the 1980 Radio Amateur's Handbook (24). Correspondingly, it has been calculated that a 7.2 pF capacitor should be used with the inductor to form the input tank circuit of a 163.5 MHz amplifier. The output tank circuit of the 163.5 MHz amplifier, shown in Figure B1b, was made to be identical to the input tank.

The points where the input lead and the gate lead should be connected to L1, for a maximum voltage gain condition, were determined experimentally as is shown in Figure B2. The maximum voltage gain occurred when the gate lead was connected to the top of the tank circuit and the input lead was connected to the center of L1. In a similar manner, the points where the drain lead and the output lead should be connected to L2, for a maximum voltage gain condition, were determined. The maximum voltage gain occurred when both the drain lead and the output lead were connected to the top of the tank circuit.

A coupling capacitor of 20 pF was chosen at the input since it has an impedance of 50Ω and helps match the input of the amplifier to the output of the preceding stage. A 20 pF coupling capacitor was also used at the output, although it seems that a smaller capacitor would probably have less effect on the tuning of the output tank.

2. Tuning

1. The neutralization components were selected experimentally so that amplifier oscillations ceased and so that voltage gain could still be achieved by proper adjustment of C(I) and C(O). It was found that the magnitude and phase angle of the neutralization impedance are important. Even though a pure inductance could be used, as is shown in Figure B1a, to stop oscillations, the amplifier could not achieve any voltage gain once the oscillations ceased. It was found that a pure resistance of 4.3kΩ was almost sufficient but that four or five turns of a fine wire wrapped around the resistor and tied to one end, as shown in Figure B1b, were necessary.
2. C(I) and C(O) were then adjusted so that maximum gain was achieved. A gain of approximately 16 db was obtained.



TAP POINTS TP1 AND TP2 ARE ADJUSTED FOR MAXIMUM VOLTAGE GAIN FROM V1 TO V2. CIRCUITRY ENCLOSED BY DASHED LINES IS THE INPUT TANK OF AMP. IN FIG. B1b WITH INPUT COUPLING CAPACITOR. INPUT Z IS 50Ω. OUTPUT Z IS COMMON-SOURCE TRANSISTOR INPUT Z.

Figure B2. Experimental setup for determining tap points (TP1 and TP2) for inductor L

XI. APPENDIX C

A. Conditions for Oscillation of the Grounded-Base Crystal Oscillator

First, a brief discussion is given on the conditions necessary for oscillations to occur in a grounded-base crystal oscillator. Then, a brief discussion is given on the characteristics of the crystal so that a more complete understanding of the circuit structure may be obtained.

The following discussion on the conditions necessary for oscillations to occur in a grounded-base oscillator is taken from pages 89-91 of the crystal oscillator book by Frerking (12). Basically, for oscillation to occur in the circuit shown in Figure C1, the phase shifts from e3 to e1, from e1 to e2 and from e2 to e3 (e1, e2 and e3 are voltages) must sum to zero, and the net voltage gain of this loop must be at least one. In more detail: First, amplification of the signal at the emitter gives e1 = (e3)(Y(fb))(Z(t)), where Z(t) is the tank impedance, which should be resistive at resonance, and where Y(fb) is the forward transfer admittance of the grounded-base transistor, which will probably have a slightly lagging phase shift at higher frequencies. Second, if C2 is fairly large compared to C1, then e2 ≈ e1 (C1/(C1 + C2)) = -e1(X2/XL), and the phase shift is negligible (although slightly leading). Finally, e3 ≈ (e2/(1 + Z(e)Y(ib))), where Z(e) is the crystal impedance and Y(ib) is the grounded-base transistor input admittance. If the crystal is at series resonance, and if the input admittance is only slightly inductive, then this simplifies to e3 ≈ e2(R(in)/(R(in) + R(e))), where R(in) is the input resistance of the transistor and R(e) is the crystal resistance. The phase shift that does tend to occur tends to compensate for the

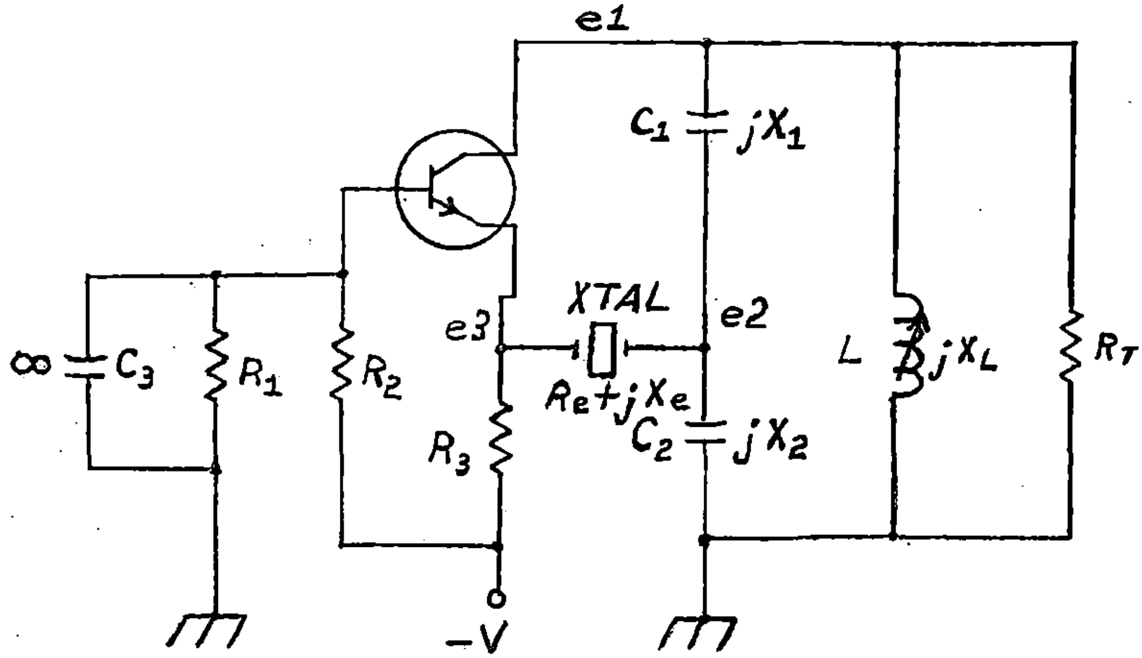


Figure C1. Schematic of a grounded-base crystal oscillator (from page 90 of Reference 12) phase lag in the transistor. These equations can thus be used as a design guide for designing crystal oscillator circuits.

Briefly, the crystal oscillator functions as follows. The frequency of oscillation adjusts itself so that the crystal presents a reactance to the circuit which satisfies the phase requirement. The oscillation then builds up to the point where nonlinearities decrease the loop gain to 1. (If the circuit is such that a loop gain greater than unity is not attainable at a frequency where the phase requirement can be met, then oscillation will not occur.) A more detailed description of the conditions necessary for oscillations to occur in a grounded-base oscillator can be found beginning on page 91 and in Appendix C of Frerking (12).

Next, a brief discussion of the characteristics of an overtone mode crystal and its use in the grounded-base crystal oscillator is given. This discussion takes material from page 30 of Frerking (12).

The reactance versus frequency curve of an overtone crystal is approximately as shown in Figure C2. If the oscillator circuit is tuned so that the crystal impedance is resistive or slightly inductive, then as can be seen in Figure C2, the oscillator can operate at a number of frequencies. According to Frerking (12), the crystal response which has the lowest resistance will dominate as oscillations build up (when power is applied) and all others will die out. Thus, if the parallel tuned circuit in Figure C1 was not present, then the fundamental frequency of the crystal would dominate. (The crystals used in the research described in this paper are designed to operate on either the 3rd mechanical overtone or on the 5th mechanical overtone (uncertain).) To prevent the crystal from oscillating on the fundamental or less than desired mechanical overtone, a tuned circuit, tuned to be resonant at the desired mechanical overtone, is used so that the zero phase shift and unity gain conditions can only occur near the desired mechanical overtone. Then, assuming that the mechanical overtone response has less crystal resistance than the nearby spurious responses, it will dominate, and the spurious responses will die out. (Information about problems of spurious responses dominating over the desired response can be found beginning on page 128 of Frerking (12).)

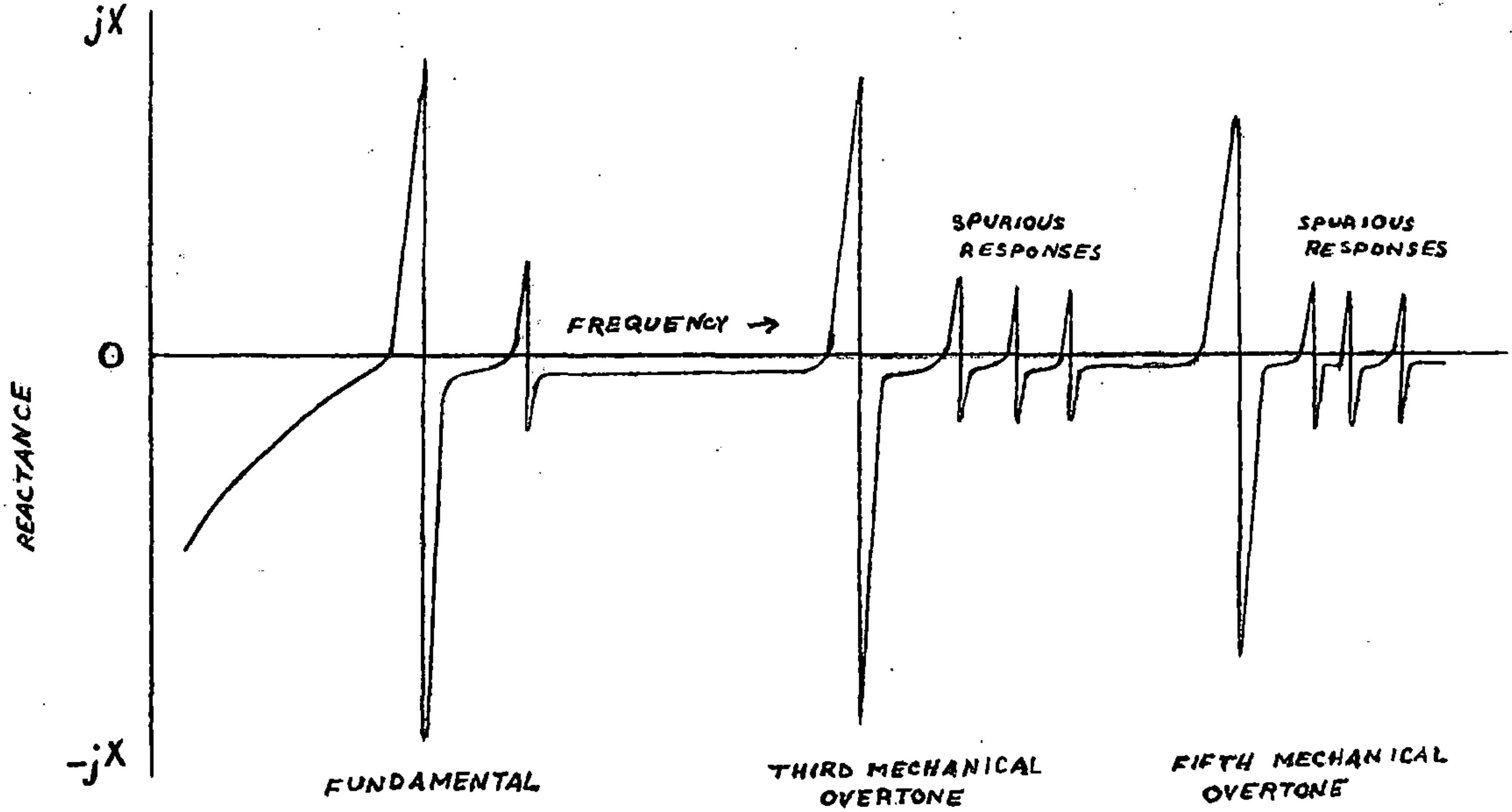


Figure C2. Overtone response of a quartz crystal (from page 30 of Reference 12)

XII. APPENDIX D

A. Construction and Tuning of a Grounded-Base Crystal Oscillator

The grounded-base crystal oscillator circuit used in the research presented in this paper was taken from page 97 of Frerking (12) and modified slightly so that the circuit in Figure D1 was the result. Only the component values were changed and not the arrangement of the components themselves. The only significant change was a 2.4kΩ (instead of a 4.7kΩ) resistor in the tank circuit of the crystal oscillator, which was selected experimentally so that the oscillator would output the desired waveshape. The other components shown in Figure D1 differ from Frerking only because these component values were more readily available.

1. Construction

Carbon composition resistors, ceramic disk capacitors and an ECG 229 transistor were used in this circuit. The crystal used is an M-tron MRH-2 designed to resonate at 53.51666 MHz. These components were placed on the printed circuit board as shown in Figure D2. Then the inductors were made using forms found in an inductor assortment pack from Radio Shack. L2 was formed by removing 6 turns or so from a 3.3 μH coil. However, L2 is not as critical as the other inductors in the circuit. L1, L3 and L4 were formed using a fine wire, such as #26, on a 6 mm diameter form. L1 was calculated to be parallel resonant with the series combination of a 5 pF and a 39 pF capacitor at 53.5 MHz. L3 was calculated to be series resonant with a 2.2 pF capacitor and L4 was calculated to be parallel resonant with a 2.2 pF capacitor, both at 160.5 MHz. L1, L3 and L4 were

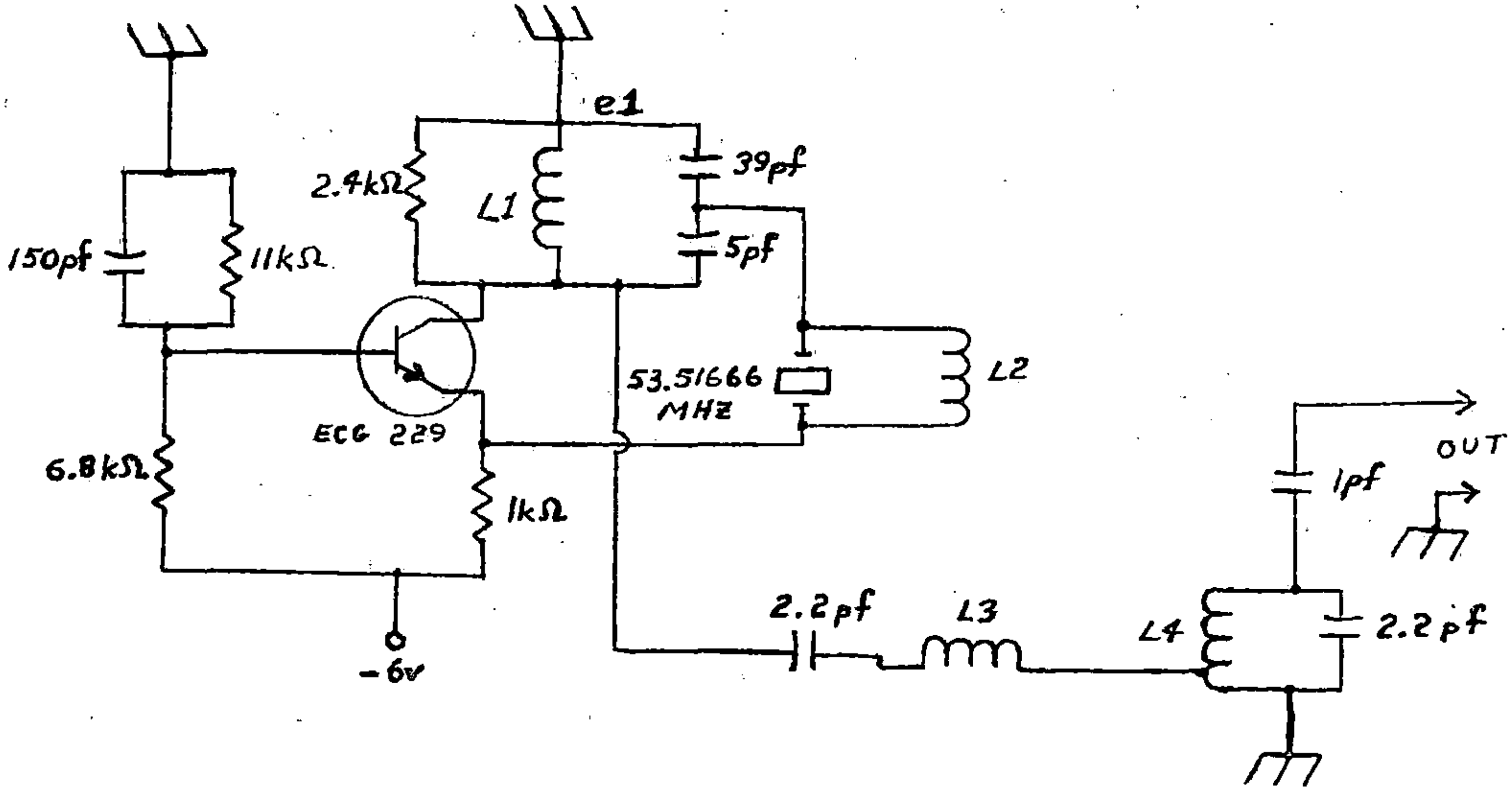


Figure Dl. 53.51666 MHz grounded-base crystal oscillator with output frequency tripler stage

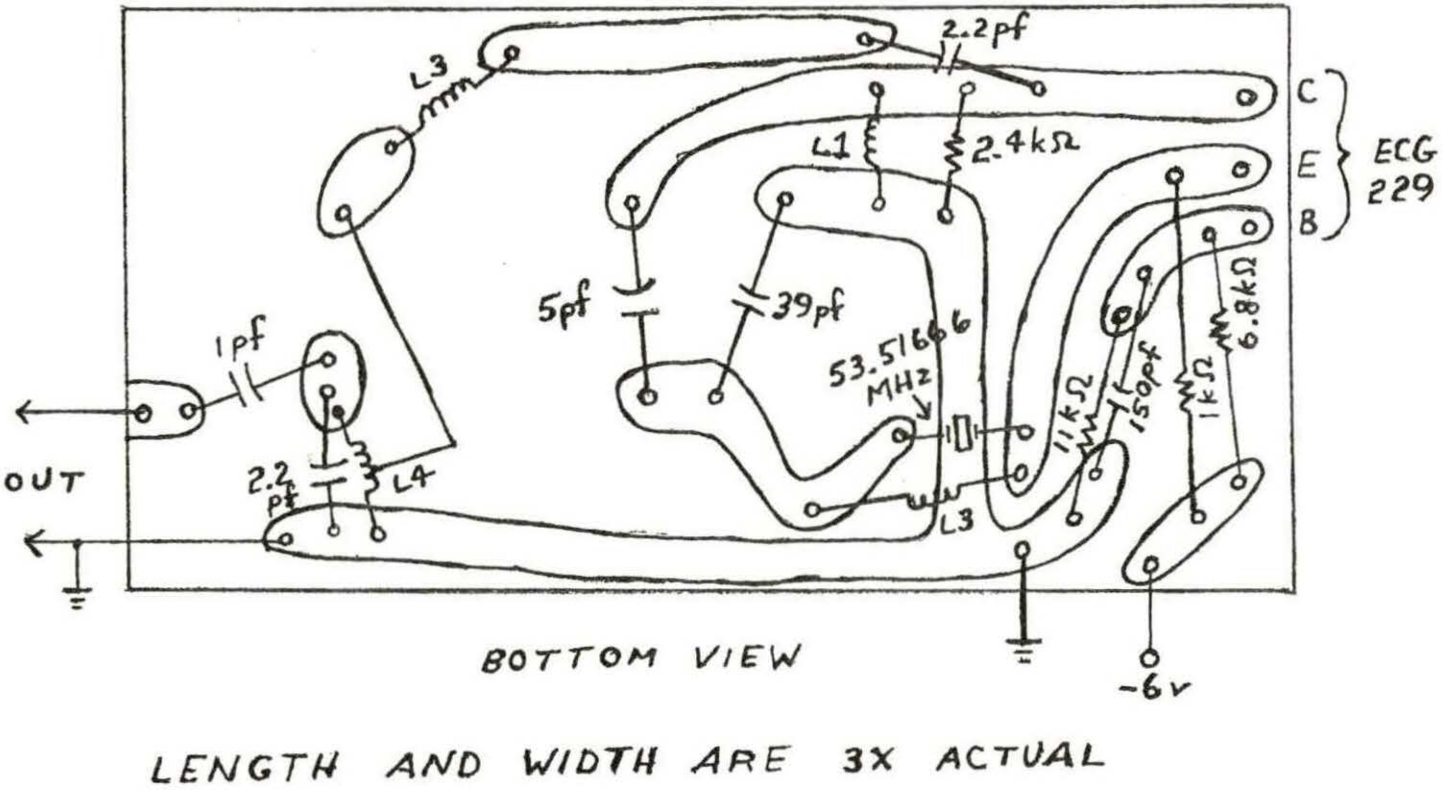


Figure D2. Approximate (enlarged) circuit board layout for circuit of Figure D1 then constructed according to the inductance formula, $L(\mu H) = (a)(a)(n)(n)/(9(a) + 10(b))$, from page 26 of the 1976 Radio Amateur's Handbook (23), where a = coil radius in inches, b = coil length in inches (for this circuit, $b \leq .25''$ was used) and n = number of turns. This formula is a close approximation for $b \geq .8a$. The tap which was made to L4 was made about two turns above ground.

These inductors were then soldered to the printed circuit board along with twisted pair wires for power supply leads and output signal leads.

2. Tuning

The signal at el was monitored while tuning L1 (L1 can be tuned by adding or removing turns to the form, or it can be made with a couple less turns than calculated, initially, and a slug can be added to the form to make tuning easier) using a Ramsey CT-70 frequency counter and a Tektronix 661 sampling oscilloscope. First, L1 was tuned so that the frequency of oscillations was approximately the crystal frequency. Then L1 was adjusted more finely so that the desired waveshape was obtained at the crystal frequency. (For the crystal oscillators used in the research presented in this paper, a distorted waveshape rich in odd harmonics was desired.) It was found that tuning the inductor was somewhat critical. While adjusting the inductor, when crystal oscillation is approached, the oscillation frequency almost jumps to near the crystal resonant frequency. (The frequency change is not gradual with change in inductance.) It was also found that the oscillator was extremely difficult to tune without the frequency counter. L1 required about ten turns to achieve the desired oscillation.

If a frequency multiplier is to be used, it can then be adjusted. For the frequency tripler stage used in the research reported in this paper, L3 and L4 can then be adjusted, as needed, to achieve the desired triple frequency output. Again, the tuning of L3 and L4 are somewhat critical. Even the position of L3 and L4 on the circuit board may be critical. L3 and L4 required about seven turns to achieve the desired triple frequency output.

XIII. APPENDIX E

A. Operation of Transmitter Address Encoding and Receiver Address Decoding Circuitry

The transmitter address encoding circuitry is shown in Figure 13. During steady-state conditions, a logic 0 is output at pin 3 of the 4021, and thus a 400 Hz signal is output at pin 4 of the 4046. When the data at pins 15, 14, 13, 4, 5, 6, 7 of the 4021 are to be transmitted, (from right to left) the reset switch must be pressed and released so that pin 12 of the 4027 becomes logic 0, and then the transmit switch must be pressed and released so that pin 15 of the 4027 becomes logic 1. On the following rising edge of the 12 Hz astable multivibrator, pin 11 of the 4011 goes to logic 0. On this falling edge, a logic 0 spike is produced at pin 2 of the monostable multivibrator which causes a 10 ms logic 1 pulse to be output at pin 3 of the monostable multivibrator. During this 10 ms logic 1 pulse, pin 3 of the 4011 goes to logic 0. When pin 3 of the 4011 goes to logic 0, pin 4 of the 4011 goes to logic 1. This logic 1 causes the data at pins 15, 14, 13, 4, 5, 6, 7 of the 4021 and a start bit to be loaded into the shift register of the 4021. Since the rightmost bit in the shift register is the start bit (logic 1), pin 4 of the 4046 becomes a 1200 Hz signal. On the rising edge of the logic 0 pulse at pin 3 of the 4027, pin 1 of the 4027 goes to logic 1 to prevent another transmit command (causes pin 15 of the 4027 to go to logic 0). On the following rising edge of the 12 Hz astable multivibrator, the address in the shift register of the 4021 is shifted one bit to the right. The start bit is lost and a logic 0 is loaded into the leftmost flipflop of the 4021 shift register. After seven more rising edges of the 12 Hz astable multivibrator, the entire address has been transmitted, and the 4021 shift register is again filled with logic 0s.

The receiver decoding circuitry is shown in Figure 14. During steady-state (after the reset switch has been pressed), the 12 Hz astable multivibrator is disabled so that no input at pin 11 of the 4021 is shifted into the 4021 shift register. When a rising edge is detected at point X1, a logic 0 spike is applied to pin 2 of the monostable multivibrator which causes it to output a 10 ms logic 1 pulse at pin 3. Thus, a 10 ms logic 0 pulse is applied at pin 13 of the 4027. If, upon the rising edge of the logic 0 pulse, logic 1 is still at pin 10 of the 4027, a logic 1 is output at pin 15 of the 4027. The 10 ms delay is used to make sure that a start bit has been seen at pin 11 of the 4021 (not just a noise spike). When pin 15 of the 4027 goes to logic 1, pin 6 of the 4023 becomes logic 0 (provided logic 0 is in the rightmost bit of the 4021 shift register, which is how it should be if the reset switch was depressed and released after the previous reception) and in turn, pin 6 of 4049-1 becomes logic 1, thus enabling the 12 Hz astable multivibrator. Pin 9 of the 4023 thus becomes the inversion of the astable multivibrator output and pin 12 of 4049-1 becomes the inversion of pin 9 of the 4023. Thus, on the first rising edge of the output of the 12 Hz astable multivibrator, the start bit is shifted into the 4021 shift register. After seven more rising edges of the 12 Hz astable multivibrator output, the start bit occupies the rightmost D-flipflop of the 4021 shift register. A logic 1 in this position causes pin 10 of 4049-1 to become logic 0. This in turn causes pin 6 of the 4023 to become logic 1 and then pin 6 of 4049-1 to become logic 0. This logic 0 then disables the 12 Hz astable multivibrator to prevent the received address from being lost. When pin 3 of the 4021 becomes logic 1, the three ICs at the lower left of Figure 14 (4049-2, 4081-1, 4081-2) decode the data bits at pins 2 and 12 (pin 12 is the most significant bit) of the 4021 and light one of the four lights shown at the right of Figure 14. After reception, whether or not pins 2 and 12 of the 4021 are logic 1, the reset switch must be pressed and released before another address can be received.