

A TELEMETRY TRANSMITTER
FOR HIGHWAY DATA COLLECTION

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by

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CHAPTER I

INTRODUCTION

In situ measurements of highway roadbed materials often present several formidable problems. Often it is impossible or undesirable to connect the measuring sensor or transducer to the data readout instruments via wire. Such is the case of measurements made in a highway roadbed. The exterior surface of the highway cannot be disturbed as it would modify its structure and thus, lead to incorrect measurements. The solution is to telemeter the measurements via radio. The radio telemetry device must be miniaturized and powered by some means whereby the device may remain undisturbed in place for a period of years.

The subject of short range telemetry is not extensively covered in the literature. Somewhat more has been written in the field of medical and biological telemetry than in industrial applications. Medical and biological radio telemetry uses have different design goals and restrictions. This is not to say that there are no common design goals between this and industrial telemetry; but that usually a particular package designed for an industrial application could more readily be adapted to a medical or biological application than vice versa.

The short range radio telemetry problem is well presented by Hoepfner [7] who points out the differences in missile and space

telemetry and industrial short range telemetry. The system inefficiencies permitted in the latter can be used to advantage in designing a simple, low cost, but yet effective telemetry system. Hoeppner breaks down the telemetry system as a transducer modulating a subcarrier oscillator which in turn modulates an RF carrier, varying its frequency or amplitude in accordance with the sub-carrier voltage signal. He goes on to suggest that the RF carrier be placed in the 88 to 108 MHz FM broadcast band. He cites the availability of high-grade radio tuners already mass-produced for the high-fidelity market which can be used to receive the telemetry signal in that band.

The use of the 88 to 108 MHz commercial FM band for short range radio telemetry is fairly common. Stark [11] chose this band to operate a radio telemetry system for measuring parameters from moving components of test automobiles. Stark's transmitter consisted of a strain gage bridge driving a differential amplifier which in turn drives a pulse duration modulator. The pulse duration modulator is a tunnel-diode coupled pair oscillator which modulates a tunnel-diode RF oscillator. The approximate RF output power is given as 50 μ W. Stark reported a typical operating time of 4 to 40 hours with this transmitter. This is about three orders of magnitude smaller than required for a highway telemetry system. In addition, the tunnel-diode transmitter often presents a critical biasing problem, but at the same time operates on relatively low

supply voltages. However, since Stark uses frequency modulation of the tunnel-diode oscillator and has physical access to the transmitter, many of these potential problems are of no consequence in the applications in which he is concerned.

Adler [4] provides information of more practical importance to the highway telemetry problem. Although he does not divulge his actual telemetry transmitter design, he does discuss the various considerations of radio telemetry measurements of temperature and strain. It should be noted that he also recommends the use of the 88 to 108 MHz commercial FM broadcast band. Adler "pots" his transmitters for protection against their environment which is not mentioned or done in the other papers. Potting the transmitter is another dimension of the problem. If properly done, it protects the transmitter from moisture, stress, and shock. There are many materials which are suitable for potting electronic circuits. Further information relating to various types of short range radio telemetry can be found by consulting the extensive bibliography in Adler's paper [4]. So the problem at hand becomes one of designing a radio telemetry transmitter which can easily interface with a variety of sensors. The general design objectives formulated at the outset of this project are:

- (1) The transmitter power consumption should be as low as possible and still retain adequate RF power output and reliable operating features.

(2) Sufficient RF energy must be radiated from the circuit itself or a very short antenna for reliable reception when the transmitter is buried up to 24 inches.

(3) The telemetry transmitter must be physically small, yet rugged enough for reliable operation in its intended environment.

(4) The design should be simple and use a minimum number of components which will contribute to a low cost per unit design.

The final design required some compromises in these objectives. However, it is felt that the best possible compromises were made and that the design attained the goals set forth.

CHAPTER II

TELEMETRY TRANSMITTER DESIGN

Power Supply Design

The method of supplying power to a buried telemetry transmitter is a key problem in the overall design. Once placed in service, the telemetry transmitter cannot be disturbed. Two methods were investigated. The first method centered on supplying power to the buried device via magnetic coupling. This method has been used on similar telemetry transmitters implanted in the human body. Kadefors, Kaiser, and Petersen [8] investigated six rectifier circuits for use with electromagnetic power transport. Their application did not demand a large separation between electromagnetic coupling coils. Their optimum rectifier/coil combination would produce, at best, about 500 millivolts at a coil separation distance of 80 mm. Thus, their results indicated that adequate power transfer into a telemetry transmitter buried approximately two feet would require a very powerful external generator and, possibly, a large pick up coil on the transmitter. Based on this data, it was decided to investigate the minimum power required by a suitable telemetry transmitter and the maximum life that dry cell batteries could be expected to give under such a power drain. If this method could prove successful, it would reduce overall system complexity, cost, and size. With this in mind, the battery power supply was pursued further.

The rechargeable nickel-cadmium cells provide the highest energy capacity (per unit volume) of the commonly available dry cells. A "D" size cell can be obtained with a 4 ampere-hour capacity. The cell voltage is 1.25 volts. The typical current drain of one of the transmitters developed was $100\ \mu\text{a}$ at 6 volts. Thus, five nickel-cadmium batteries would operate the transmitter continuously for approximately 40,000 hours or 4.56 years. The battery pack would have to be approximately 2.5" X 3.75". The size of the pack could be reduced by designing the transmitter to operate from a lower voltage. This would have resulted in a good design in terms of physical size and lifetime. However, these cells and others (zinc-carbon, alkaline, and mercury) operate at sharply reduced capacity as the temperature drops below 20°F . Since the telemetry transmitter must operate continuously during all seasons, the lifetime would be reduced and erratic transmitter operation may result in the winter months.

Recently a new battery became available on the market using Lithium [6]. The cell voltage is 3 volts and rated at 8.0 ampere-hour capacity for a size "D" cell. Four of these cells were purchased from Chromalloy Electronics Division. This manufacturer claims that the cells can be stored for up to 10 years (70°F) and still operate with a maximum loss of capacity of 15%. Additionally, they claim the cell will operate at -20°F at 90% of capacity and at -40°F at 60% of capacity! The rated capacity is calculated on

moderate loads (i.e. 100 ma for one "D" cell) to the point that the terminal voltage drops to 2.0 volts. However, for extremely light loads, the battery capacity rating may be extended by about 25%.¹

The one temporary disadvantage in the use of the Lithium cells for the transmitter power supply is the high cost per cell. Presently the "D" size cell costs \$11.48 in quantities of less than 100. Thus the battery supply for each transmitter is nearly \$23.00. However, the price is expected to drop as production increases and more companies begin manufacturing the batteries [6]. Eventually it is predicted that they will become competitive in price with conventional cells. Thus, the present high cost of the cells did not deter incorporating them in the design of the telemetry transmitters.

Transmitter Design

It was initially determined that the transmitter would consist of an oscillator stage and some type of modulator. A junction FET was chosen for the oscillator because of the inherently high input impedance and low input/output capacitance of the device. Additionally, the use of an FET allows proper biasing with fewer resistors than a BJT; thus, saving power consumption and reducing the overall transmitter size. It was also decided that the transmitter should be crystal controlled for best frequency stability

¹Private conversation with Dr. Anthony Fraioli, Plessey Inc.

and reliable oscillation. However, the ability of a crystal oscillator to operate properly over the range of stresses encountered while embedded in a highway roadbed was questionable. Crystal specifications of a major U.S. crystal manufacturer were consulted and it was judged that indeed a high quality crystal, manufactured to military specifications, should survive the rigors of encapsulation and burial in a roadbed. Therefore, the first oscillator designed and built was an FET crystal oscillator.

The oscillator circuit employed was commonly known as a tuned-grid tuned-plate in the vacuum tube oscillator days. In this case, the FET replaced the triode tube and the crystal is the tuned circuit in the gate circuit. The basic oscillator circuit is shown in Fig. 2-1. Ordinarily the source is grounded directly. Here, the source is self-biased with a $15\text{ K}\Omega$ resistor in order to achieve a low drain current. The source is bypassed to ground with the 0.001 microfarad capacitor to provide the necessary low impedance source return at the oscillating frequency. The parallel tuned network in the drain circuit is resonate at the crystal frequency. Since a 44.1 Mhz crystal is necessarily an overtone type, the tuned drain circuit insures that the crystal is excited at the proper overtone frequency. The FET chosen for this design was a Motorola type MPF102. The MPF102 specifications are well suited for this application and is readily available for less than one dollar.

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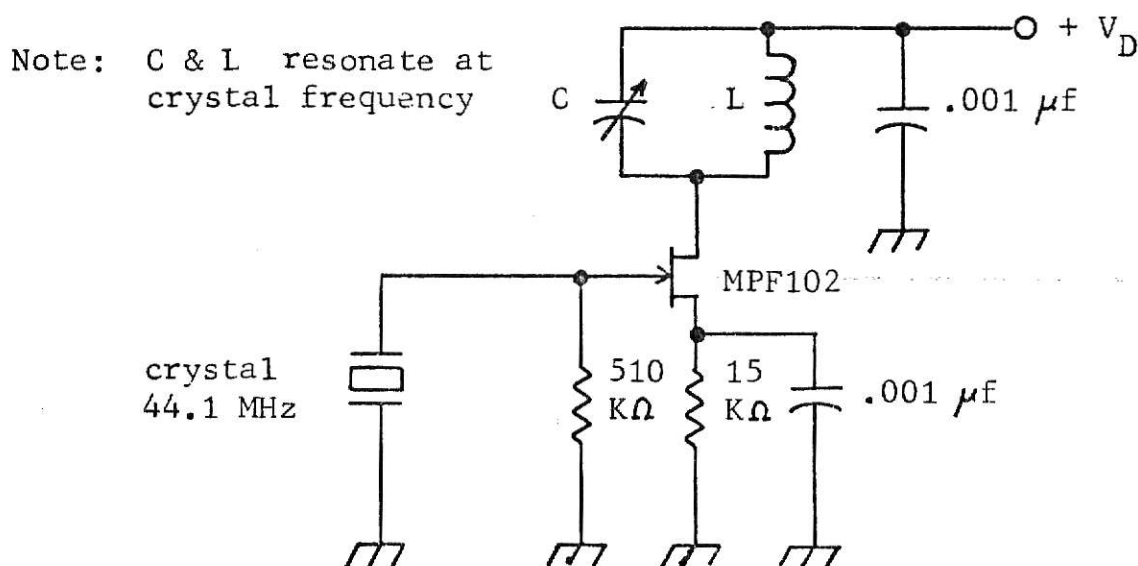


Fig. 2-1 Basic FET crystal oscillator circuit.

When the oscillator was built and determined to be functioning properly, the total current drain was measured and found to be approximately 180 microamperes when operated from a 9 volt battery. This was determined by inserting a Triplet (Model 930) VOM in series with the supply voltage. Later, a more accurate method of current measurement was employed. The oscillator was checked for range by listening to the second harmonic (88.2 Mhz) on a Sansui 800 FM stereo receiver connected to a simple dipole antenna. Adequate signal strength was received ("full quieting" the receiver) up to approximately 50 feet between receiver and oscillator. These initial crude tests indicated that it was indeed feasible for a simple FET oscillator to operate on very low power and still radiate sufficient energy from the circuit wiring for short range

reception. However, before proceeding with design refinements, the oscillator gate circuit was modified for direct frequency modulation (FM) of the oscillator. Again, no elaborate testing was accomplished. It was deemed necessary only to be able to FM the oscillator while maintaining the center frequency stability afforded by the crystal. The circuit of Fig. 2-2 met these requirements.

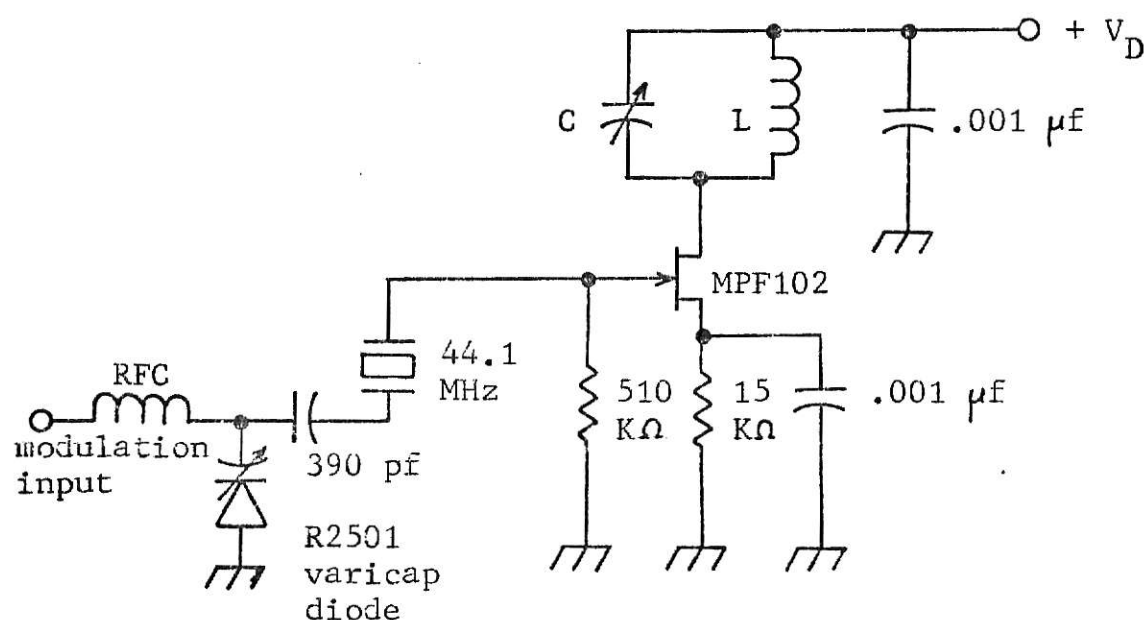


Fig. 2-2 Frequency modulated crystal oscillator circuit.

The next oscillator design was undertaken with the objectives of further reducing power consumption and operation in the commercial FM broadcast band. Two overtone crystals in miniature holders were purchased and connected to the original oscillator circuit (Fig. 2-1) with appropriate changes in the tank circuit.

The circuit refused to oscillate. Substituting different FET's and crystals was tried without success. It was then decided to try another circuit with the crystal oscillating in a series resonant mode. The circuit shown in Fig. 2-3 was constructed. It is basically a Hartley Oscillator with the crystal placed in series with the feedback path to the source. The source must be at a high impedance at the crystal frequency. The 15 K Ω source bias resistor was adequate, thus eliminating any RF choke requirement in the source circuit.

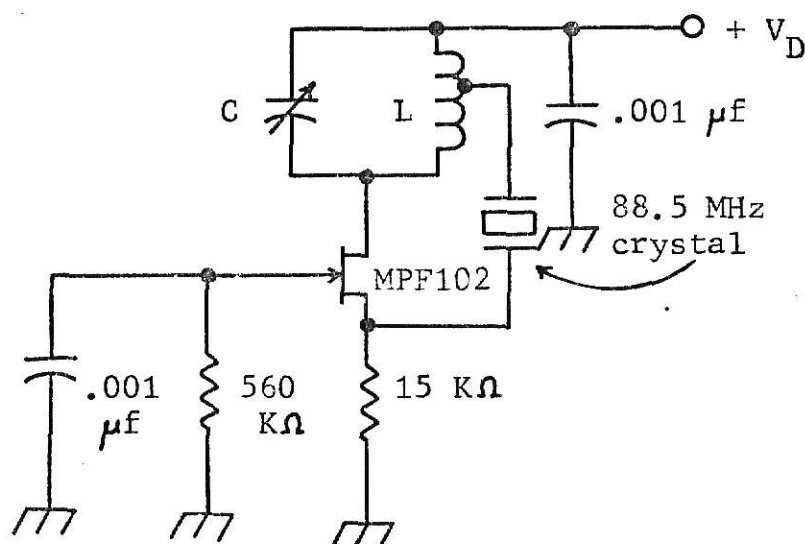


Fig. 2-3 Modified Hartley Oscillator with series crystal.

The circuit oscillated, but not necessarily at the crystal frequency as the tuned drain circuit determined the oscillator frequency. A bypass capacitor was tapped down on the source resistor to introduce a small amount of negative feedback in an

effort to prevent the parasitic oscillations, but to no avail. A check of the crystal holder specifications revealed that the holder capacitance was about 7pf! At 88.5 Mhz this is approximately 300 ohms of capacitive reactance. This capacitance apparently provided enough feedback at the proper phase angle around the crystal element to cause oscillation. A 7pf capacitor was substituted for the crystal in the circuit. Only a slight change from the previous oscillator frequency confirmed that this was indeed the case.

The Radio Handbook [10] recommends placing an inductance across the crystal to resonate with the holder capacitance at the crystal frequency. This would essentially result in a low Q parallel tuned circuit in parallel with a very high Q series tuned circuit (the crystal). This was not tried as previous experience with this technique at VHF and UHF has shown it to be quite "tricky" in adjustment. Also another component would have been required, thus increasing overall cost and size. Instead, a slightly different circuit as shown in Fig. 2-4 was tried. Since the circuit would readily oscillate with the crystal holder capacitance in the feedback path, it was replaced with a 10pf capacitor and the crystal was placed in the gate circuit. The gate impedance to ground at the oscillator frequency must be kept low in order for oscillation to occur. It was reasoned that the crystal operating in a series resonant mode would provide this low impedance path at

only the desired crystal frequency. Hopefully, the holder capacitive reactance would not cause instabilities. The crystal manufacturer stated that the maximum resistance of the crystal (at resonance) is 60 ohms while the holder capacitive reactance is approximately 300 ohms. This "brute force" method proved successful when the oscillator was constructed. The oscillator was completely stable, oscillating at only the crystal frequency.

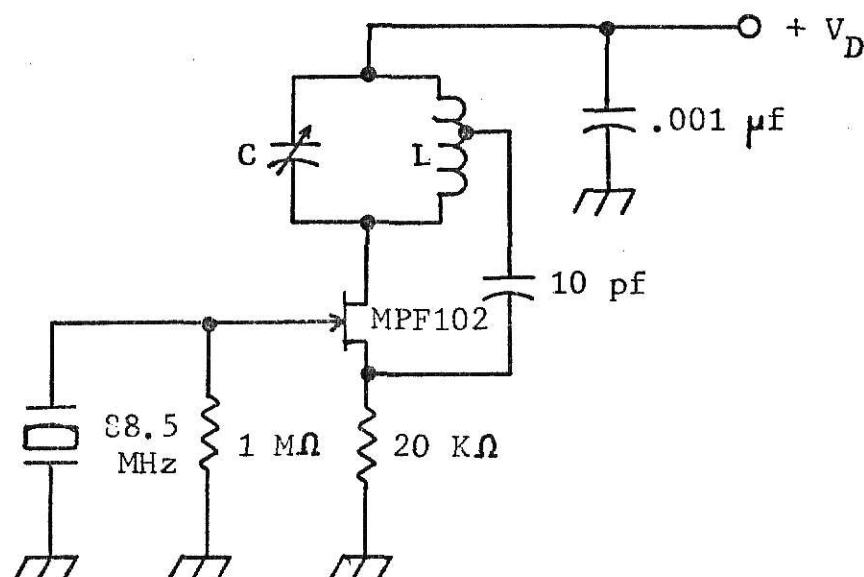


Fig. 2-4 Crystal controlled VHF oscillator circuit.

Frequency modulation of the oscillator was attempted using the circuit shown in Fig. 2-5. This circuit modulated properly as long as the capacitor between the gate and the "vari-cap" diode is chosen with care. If it is too large, the crystal loses control of the oscillator frequency. If it is too small, the frequency deviation will be reduced. The value shown was

experimentally determined to be optimum for the particular "vari-cap" diode used, the modulation applied (0 to 6 volts square wave), and for reception on a standard FM broadcast receiver.

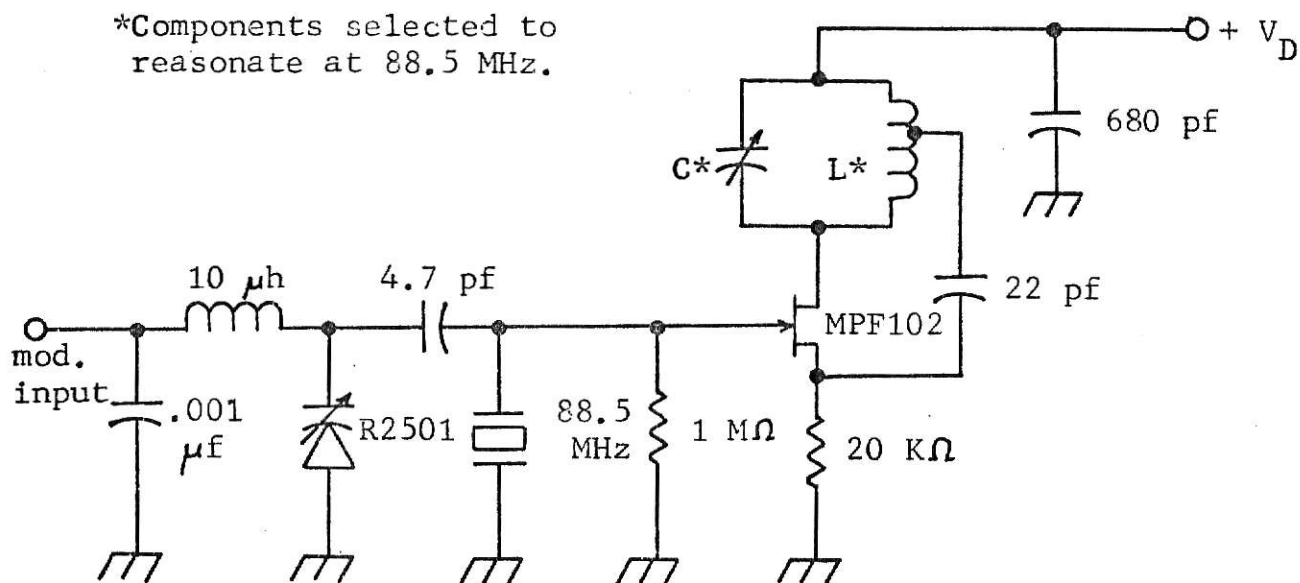


Fig. 2-5 Crystal oscillator which was successfully frequency modulated.

Modulator Design

At this point, a modulator design was required to interface the sensor with the transmitter. The Signetics NE566 voltage controlled oscillator (VCO) seemed to satisfy the requirements. It had excellent linearity, low temperature drift, and unfortunately,

a relatively high power consumption. The COSMOS family of integrated circuits was investigated. The basic COSMOS circuit element is shown in Fig. 2-6. The insulated gate provides extremely high input impedances (approximately 1000 Megohms!). Current flows from V_{DD} to V_{SS} essentially only during a transition from one logic state to another. Otherwise, one of the FET's is always cut-off, blocking all but leakage currents from flowing. The RCA COSMOS IC's include a phase locked loop (PLL) which contains a voltage controlled oscillator (VCO). The CD4046AE PLL was purchased and tested to determine its suitability for this application.

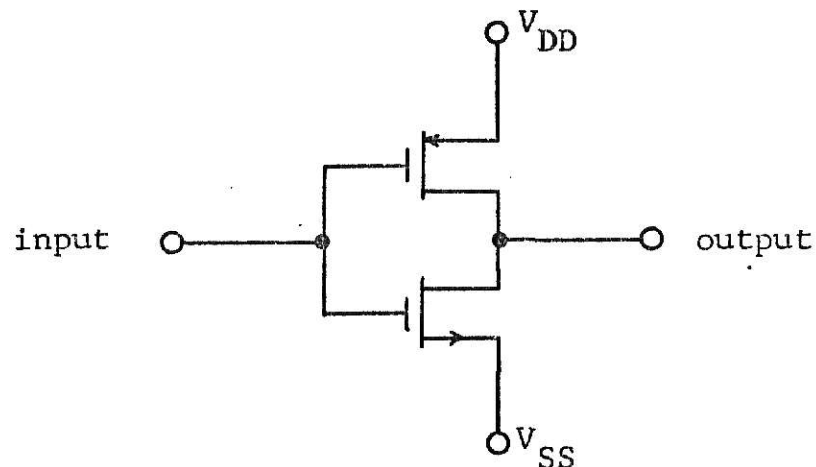


Fig. 2-6 Basic COSMOS circuit element.

The VCO in the CD4046AE has many desirable features for this application. The supply voltage can be anything from 0.5 to 15 volts, although 5 to 15 volts is the recommended range. The VCO output is a square wave between zero volts and the supply voltage. Linearity of the VCO is typically 1%. Temperature stability is

rated at 500 ppm/ $^{\circ}$ C. The specification sheet further claims a typical power drain of 100 microamperes at $V_{DD} = 6$ volts at a frequency of 10 kHz [1].

The CD4046AE was connected to a test circuit as shown in Fig. 2-7 to evaluate the performance of each unit before using them as telemetry modulators. The data book [1] provided the following formula for selecting R_1 and C_1 for a desired VCO center frequency:

$$f_o = \frac{K}{R_1 C_1} \quad \text{where: } K \text{ is a constant dependent upon the supply voltage}$$

f_o is center frequency in kHz

R_1 is in $K\Omega$

C_1 is in μf .

K for the case of $V_{supply} = 6$ Volts was given as 0.57. The data sheet specified that $R_1 \geq 10 K\Omega$ and $C_1 \geq 50pf$ as additional circuit constraints. It was reasoned that C_1 should be chosen small, thereby forcing R_1 large for a desired center frequency. This minimizes the energy required to charge C_1 (through R_1) during each cycle and consequently, minimizes the average current drain. A center frequency of 1000 Hz was chosen and C_1 chosen to be 200 pf.

$$\text{Thus: } R_1 = \frac{0.57}{(1.0)(0.0002)} = 2,850 K\Omega \text{ or } 2.85 M\Omega$$

This formula was found to indeed be only an approximation. The actual value used was 5.6 $M\Omega$ which produced an f_o of approximately 800 Hz. Current drain was measured at 20 microamperes, typically,

with the VCO input tied to V_{DD} . One unit tested drew an almost unbelievable 13 microamperes. There was also some variation of output frequencies between different units tested under identical conditions. Fig. 2-8 shows a plot made of VCO input voltage versus output frequency for eight units. Results show some difference between units, but each exhibited good linearity over a large range of VCO input voltage. These variations between units, coupled with component tolerances, point out the fact that each telemetry transmitter with sensor would have to be individually calibrated before being placed into service.

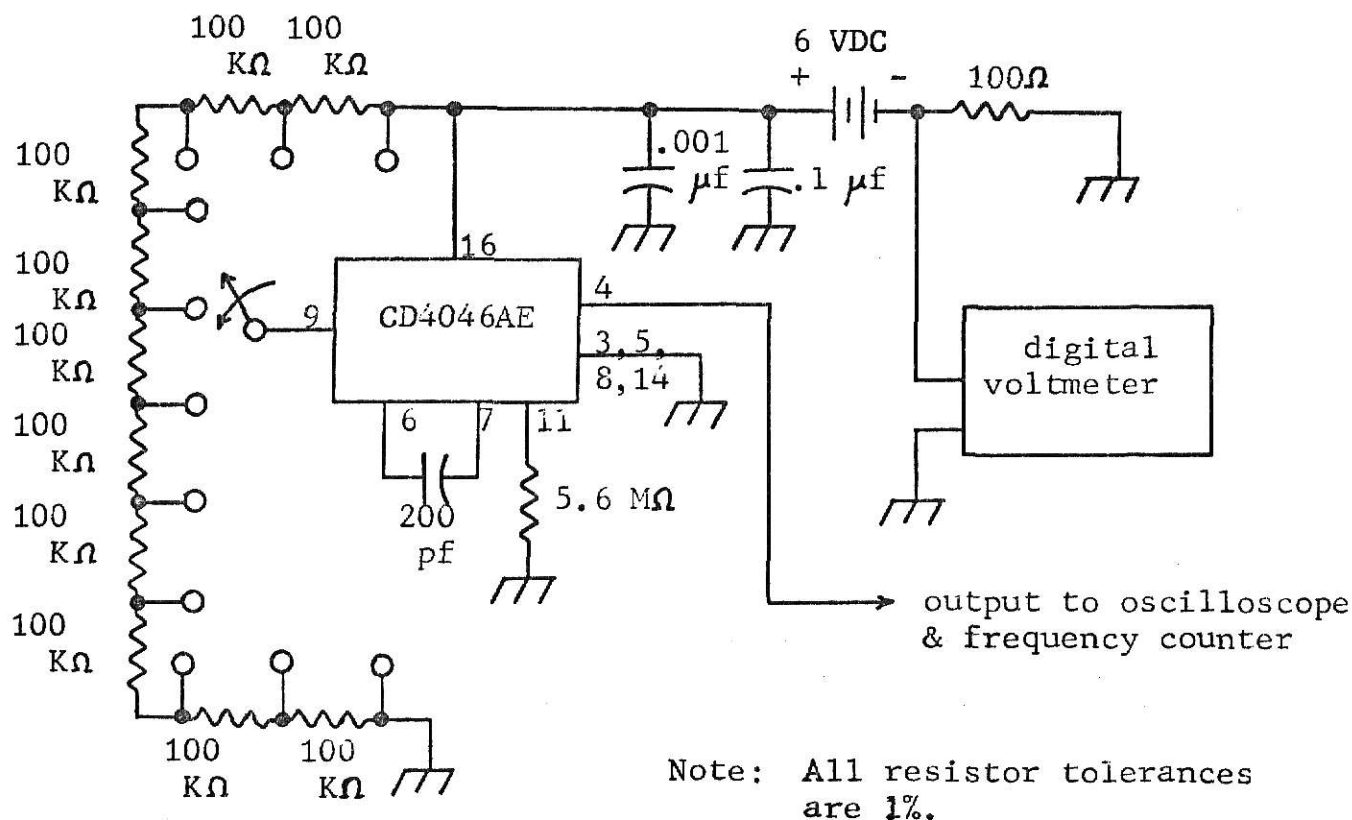
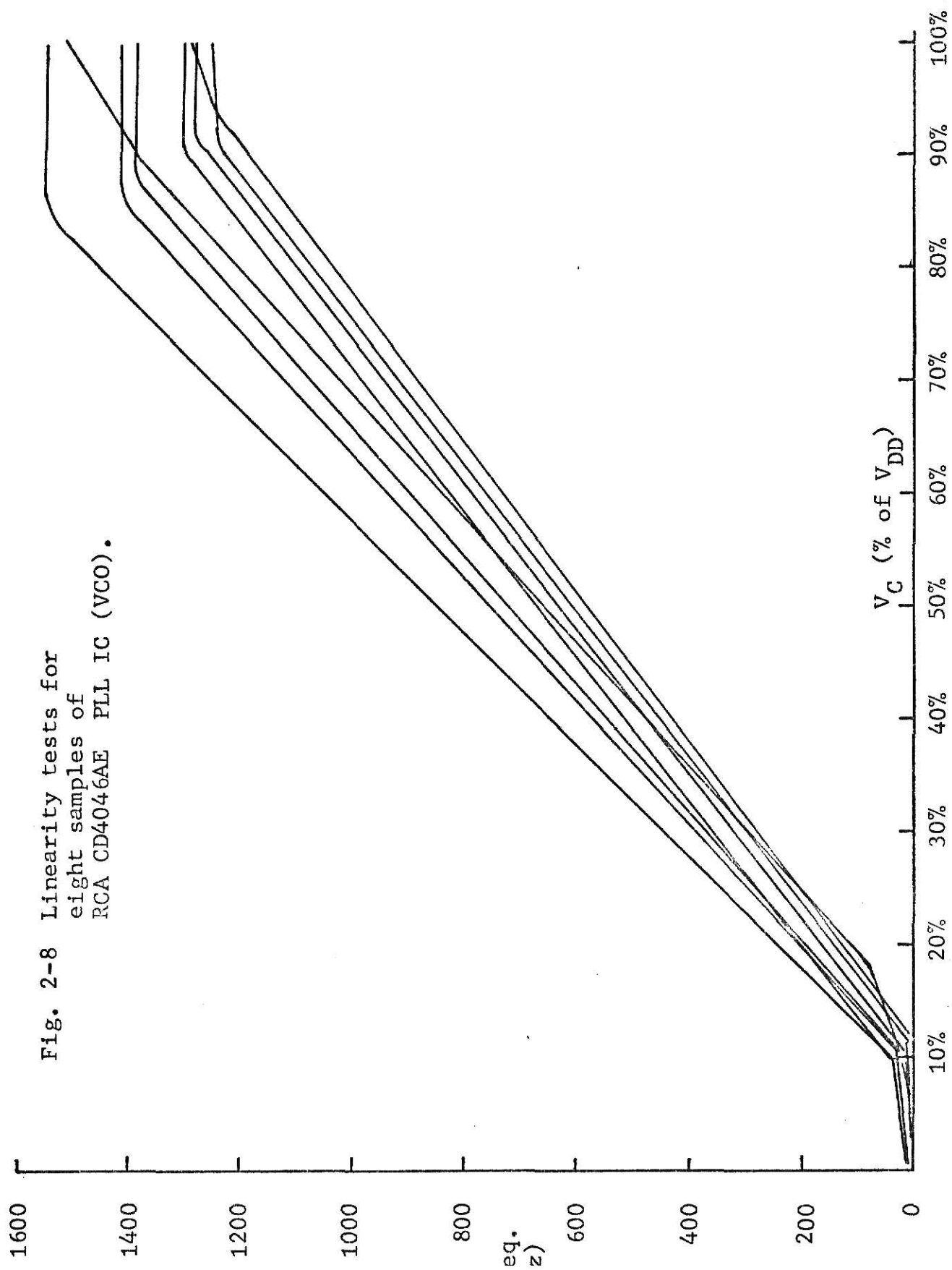


Fig. 2-7 COSMOS CD4046AE VCO test circuit.

Fig. 2-8 Linearity tests for
eight samples of
RCA CD4046AE PLL IC (VCO).



Final Design Considerations

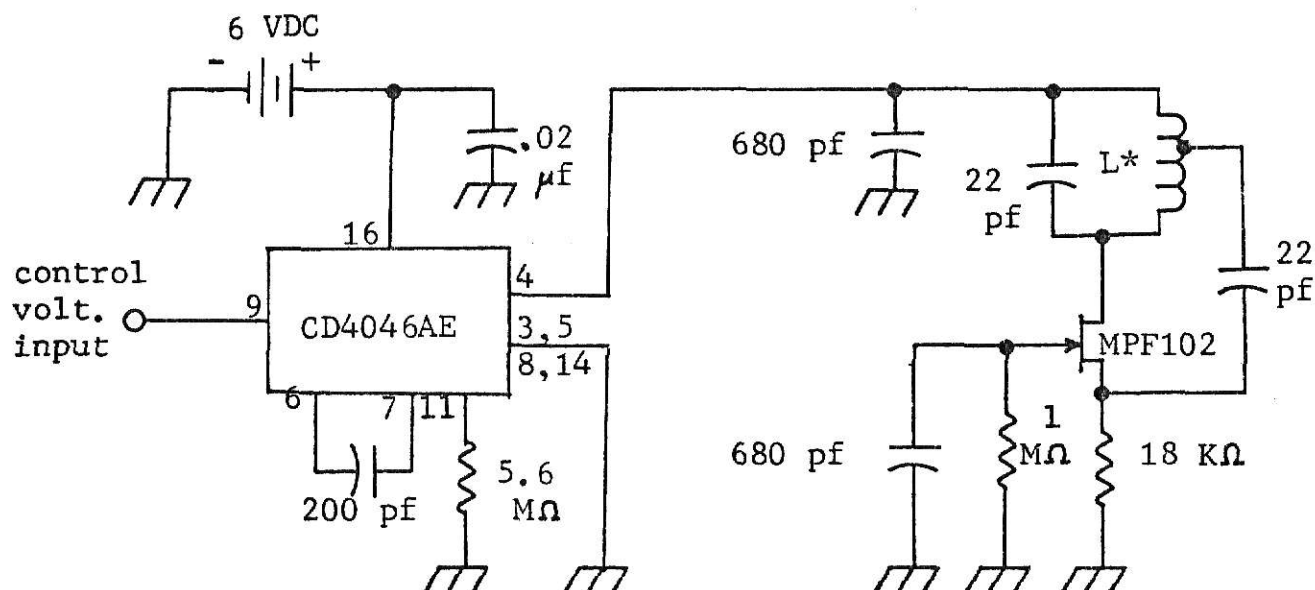
The CD4046AE VCO was now incorporated into the design of the telemetry transmitter. The first transmitter built used the VCO output to frequency modulate the crystal oscillator as shown in Fig. 2-9. This transmitter was powered for the first time by the new Lithium batteries. The VCO input was connected to V_{DD} rather than a sensor. This particular transmitter was used to study the effects of temperature on circuit and battery reliability and to conduct relative signal strength tests while the transmitter was enclosed in a small concrete box. Current drain was measured at $140 \mu\text{a}$ enabling the transmitter to operate for 57,142 hours or about $6\frac{1}{2}$ years on two 3-volt Lithium "D" cells! The figure becomes a remarkable 8.15 years if one applies the 25% longer life factor at low current drains mentioned previously. However, this does not account for reduced battery life due to operation at temperatures of 0°F and below. This effect can be estimated by assuming that part of the time the transmitters will be operated at a temperature of -20°F (winter months in northern states). We can assume worst case by using the manufacturers rating of 90% full capacity at -20°F for the entire lifetime. This would give a new lifetime figure of 7.34 years.

The breadboarded transmitter of Fig. 2-9 performed as designed. However, during the course of designing and testing this transmitter, it became apparent that perhaps there were overriding

signal could easily be received with a sensitive Hi-Fi type FM receiver. Thus it seemed that the telemetry transmitter could be redesigned for non-crystal control. This was a relatively easy task based on earlier experience in preventing self-oscillation in the crystal controlled transmitter. Without the requirement of a crystal, the transmitter could be designed with fewer components and lower cost. The crystal savings alone was \$8.15.

The new transmitter was designed along the same philosophy as the crystal controlled version. The transmitter frequency was now controlled entirely by the tuned drain circuit in the familiar Hartley Oscillator configuration. The drain tank circuit was designed with a relatively high capacitance to inductance ratio. This reduces the effect of stray capacitance on the oscillator frequency and permits the use of a physically small inductor which is required in the interests of miniaturization. It was also decided to make another major design change. The transmitter was pulse modulated directly with the output of the VCO modulator, rather than frequency modulated. The VCO output can supply 1 ma [1]. Since the oscillator consumes about 140 μ a, it is well within the VCO output rating. The transmitter average power consumption is cut to one-half since it is only on half of the time. This results in a total transmitter power drain of 85 to 90 microamperes at a supply voltage of 6 volts. This is the principle advantage of this transmitter and should allow transmitter life

to approach the shelf life of the Lithium batteries. The final design schematic is shown in Fig. 2-10.



*L = 6 turns ($f \approx 90$ MHz) or 5 turns ($f \approx 100$ MHz) of no. 20 AWG enameled wire close wound on 3/16" diameter form; tap at 2 turns from low impedance end.

Fig. 2-10 Pulse modulated telemetry transmitter.

Since the transmitter is no longer frequency modulated, the commonly available home FM receiver is not well suited for receiving the pulse modulated signal. However, it turns out that the transmitter's carrier frequency is dependent upon the oscillator supply voltage (which is changing from 0 to 6 volts at the modulation frequency) and, therefore, the transmitter exhibits a degree of frequency modulation which can be easily detected in an

FM receiver. Ideally the receiver should have an AM detector and an IF bandwidth of twice the highest frequency of the modulator (see Appendix). Since none was available and funds for this project were limited, an FM broadcast receiver was used in evaluating the telemetry transmitter's performance. In addition to having the wrong detector, the receiver has a nominal bandwidth of 200 kHz. Thus the transmitter range measurements presented later are somewhat lower than would be expected with a properly designed receiver.

CHAPTER III

TEMPERATURE AND MOISTURE SENSORS

The thermistor is probably the most suitable temperature sensor for roadbed measurements. It can be fabricated in a variety of ways and readily adapted to almost any application. Thermistor elements are semiconductors generally made by sintering combinations of metallic oxides. Manganese and nickel oxides are most commonly used for high resistance thermistors for operation at high temperatures; the further addition of cobalt oxide produces thermistors with lower resistances and for operation at low temperatures. Information on specific ingredients to produce a desired thermistor characteristic is difficult to obtain since the manufacturers of thermistors and thermistor materials regard this as proprietary information.

The thermistor is characterized by a non-linear relationship between temperature and resistance. The exponential behavior is described by the thermistor equation:

$$R_T = R_o e^{\beta(1/T - 1/T_o)}$$

where: R_T is the resistance of the thermistor
 R_o is the resistance of the thermistor
 at temperature T_o
 T is temperature in degrees Kelvin
 T_o is a reference temperature in degrees
 Kelvin (usually taken as room temperature)
 β is a constant

The constant β depends upon the thermistor material and can be changed by altering the thermistor composition. It is typically between 2000 and 4000° K. Thus, by manipulating the values of β and R_0 , a thermistor can be obtained for almost any temperature sensing application below about 300° C.

The thermistor sensor used in this project was fabricated with thick film techniques. The thermistor paste was first silk screen printed onto an alumina substrate in the desired geometry. Then it was fired and when cool, a protective glaze was silk screen printed and then fired over the thermistor layer. Conductor contacts to the thermistor element were also provided on the alumina substrate. The finished sensor is diagrammed in Fig. 3-1.

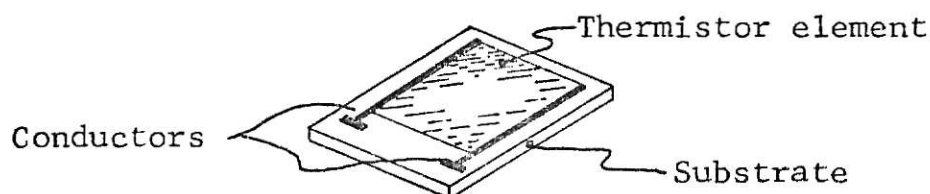


Fig. 3-1 Thermistor temperature sensor.

The alumina substrate has good heat conductivity for an electrical insulator. Typically it is about 225 BTU·in/hr·ft²·°F. Thus, one could expect fairly rapid response from the thermistor to changes in temperature.

The thermistor resistance for several values of temperature for a typical thick film sensor are tabulated as follows:

Temperature ($^{\circ}\text{K}$)	273	289	298	303	313	323	333	343
Resistance ($\text{M}\Omega$)	13.0	7.6	6.0	5.2	3.9	3.0	2.2	1.7

If we select the resistance values at 273°K and 298°K , we can calculate β for this particular thermistor as follows:

$$R_T = R_O e^{\beta(1/T - 1/T_O)}$$

$$6 = 13e^{\beta(1/298 - 1/273)}$$

$$\ln 6 = \ln 13 + \beta(1/298 - 1/273)$$

$$\ln 6 = \ln 13 - \beta(3.073 \times 10^{-4})$$

$$\beta = \frac{\ln 13 - \ln 6}{3.073 \times 10^{-4}}$$

$$\beta = 2518^{\circ} \text{K}.$$

Practical moisture sensors take many forms. Each seems to have a number of disadvantages which render them unsuitable for certain applications. Research is presently being conducted to overcome these problems. One method of determining soil moisture contents involves a capacitive sensor. Basically this method involves the measurement of the dielectric constant of material in the electric field of a capacitor. This method is one of the simplest, fastest, and cheapest methods of determining soil moisture. However, its chief disadvantages appear to be errors introduced by variations in soil dielectric constant, particle size, density, and ionized salts [3].

The capacitive moisture sensor produces a change in capacitance due to the moisture content of the soil. Oven dried soil has a dielectric constant of approximately 2.6 whereas the

dielectric constant of water is about 80. Thus the sensor capacitance increases directly with the moisture content of the soil. The exact relationship between sensor capacitance and soil moisture is complex and depends upon many factors peculiar to the particular type of sensor.

A practical capacitor moisture sensor developed at Kansas State University is shown in Fig. 3-2.

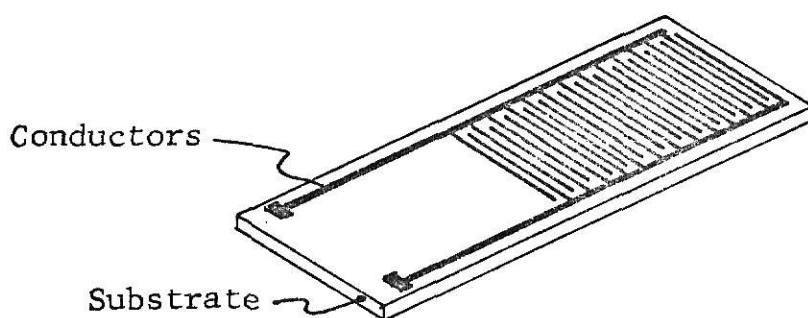


Fig. 3-2 Microcomb capacitor moisture sensor.

The sensor is in the form of a microcomb capacitor fabricated with thick film techniques. The conductors are printed onto an alumina (Al_2O_3) substrate. The conductors are composed of silver-palladium which is applied to the substrate in paste form by printing through the microcomb pattern on a silk screen. The printed substrate is then fired in an oven to produce adhesion between the conductor and the substrate. Finally a layer of glass (Cermet Resistor Overglaze, Electro-Science Laboratories No. 4770-B-Clear) is silk screened onto the substrate and then fired. The layer of glass protects the

microcomb capacitor from shorts or resistive bridges when it is implanted in soil.

A typical sensor was built with a line width of 0.007 inches and a line spacing of 0.007 inches. This sensor gave a nominal capacitance of 55 pf in air, 75 pf in dry soil, and 750 pf in water. This is a fairly wide range of values although it is not as wide as one would expect from the differences in dielectric constants. This is due to the protective glass coating. Fig. 3-3 shows a partial cross-section of the sensor. Note that a portion of the electric field between capacitor plates passes only through the glass or the substrate and thus, is not subject to external dielectric changes.

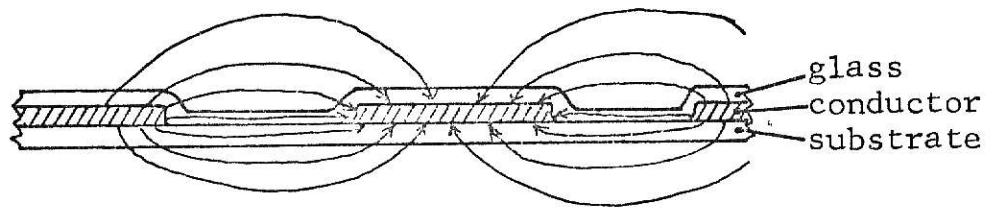


Fig. 3-3 Cross-section of a capacitive moisture sensor.

In general the sensor can be represented as an ideal capacitor in parallel with a resistor. The resistor represents a loss which occurs in the dielectric. If the capacitive sensor is used in an oscillator circuit, the shunt resistance as well as the capacitance

will determine the frequency. In most cases, the shunt resistance is much larger than the capacitive reactance.

Since a portion of the dielectric is moist soil, what effect do ions in the water have upon sensor accuracy? Monfore [9] states that the addition of 0.1 to 0.7 mole of sodium chloride has little effect on the dielectric constant of water. Since he does not state the quantity of water used, the concentration of NaCl is uncertain. Tests were conducted to ascertain the effects of NaCl in various concentrations. The sensor was used as the frequency determining element of a voltage controlled multi-vibrator (VCM). The test circuit is shown in Fig. 3-4.

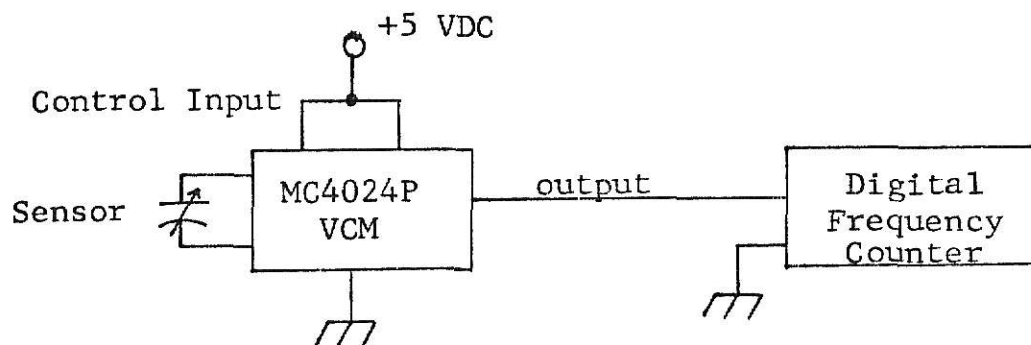


Fig. 3-4 Moisture sensor test oscillator.

The VCM frequency was 115 kHz when the sensor was placed in distilled water. The frequency was carefully measured as the percent of sodium chloride in solution was increased. The results are in Fig. 3-5.

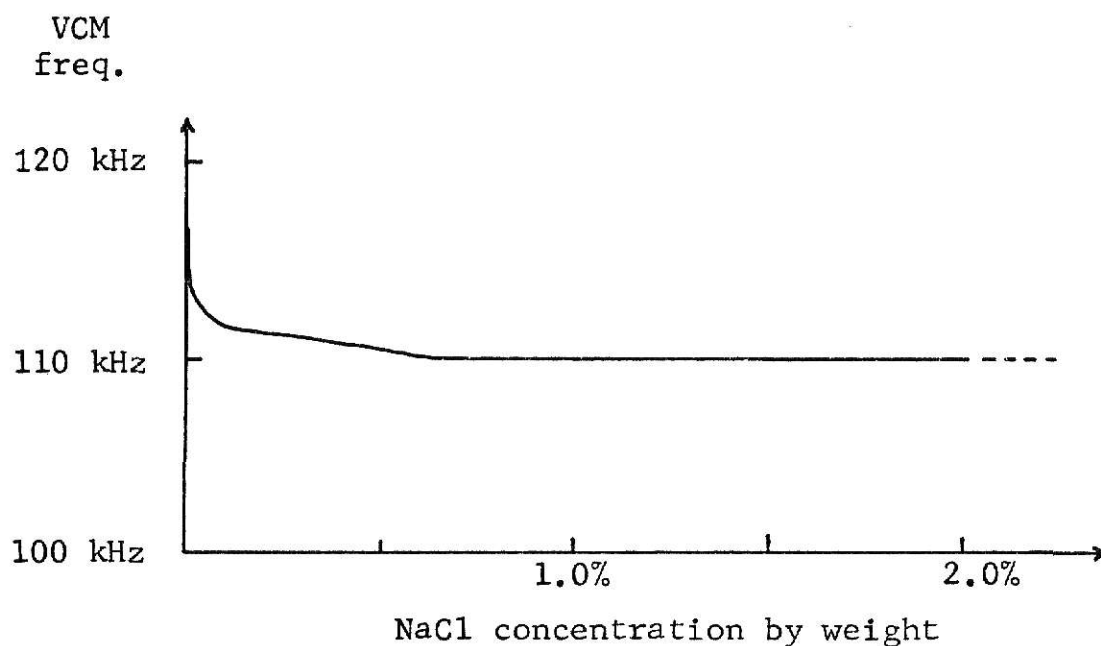


Fig. 3-5 VCM frequency vs. sodium chloride concentration.

The sensor was also tried on "de-ionized" water and a VCM frequency of 365 kHz was recorded. Thus, from this test one could conclude that the dielectric properties of water are affected a great deal at very low ion concentration levels (below 0.05%) and relatively little at higher concentrations.

CHAPTER IV

EXPERIMENTAL RESULTS

The pulse modulated telemetry transmitter shown in Fig. 2-10 was selected for testing with some experimental moisture and temperature sensors which are being developed at Kansas State University for in-situ highway measurements. These tests are designed only to show the applicability of using the telemetry transmitter as a means of extracting the sensor measurement. The testing also indicates calibration procedures which would be applicable to other types of sensors.

Since several transmitters were required for demonstration and testing purposes, a printed circuit board was fabricated for the pulse modulated transmitter. The first step was to lay out the circuit in planar form at a scale of five times normal on a piece of paper. Actually, the circuit was laid out with two physical designs to allow for flexibility in packaging. One circuit board was approximately $2\frac{1}{2}$ " X $\frac{3}{4}$ " and the other $1\frac{1}{8}$ " X $1\frac{1}{2}$ ". After the layout designs were finalized, the conductor pattern was cut into a sheet of rubylith. The rubylith was removed from the plastic backing on all areas which required copper to be etched away on the copper clad board. The rubylith was then photo-reduced to normal size using Kodalith Ortho film (type 3) with a camera lens f-stop of 16 and exposed for approximately four seconds.

The negatives were then developed, washed, and dried. Fig. 4-1 shows the conductor pattern on the negative.

The copper clad board was a high grade glass epoxy board recommended for use on high reliability applications. The copper surface was thoroughly scrubbed to remove any oxide, oils, fingerprints, ect., before coating with a photoresist solution. Two coats of photoresist were sprayed on the copper (the board was dried between coats) to insure that no "pin-holes" were left. The board was then thoroughly dried to harden the photoresist layer.

The conductor pattern negatives were placed over the sensitized board and held flat by a glass plate. The boards were exposed for eight minutes with a 250-watt ultraviolet lamp placed 12 inches above the board. The board was developed in Kodak Photo Resist Developer for 60 seconds, washed in cold water, and then dried with a hot air gun. The pattern was just barely visible at this point indicating the board had been properly exposed and developed.

The board was etched in a solution of ferric chloride to remove the unwanted copper. The speed of the etching process was increased by heating the solution to about 180° F and agitating the solution over the board by bubbling air from the bottom of the solution. This technique produced a fully etched 3" X 6" board in about six minutes. The board was then washed, dried, and examined for evidence of incomplete etching and "pin-holes".

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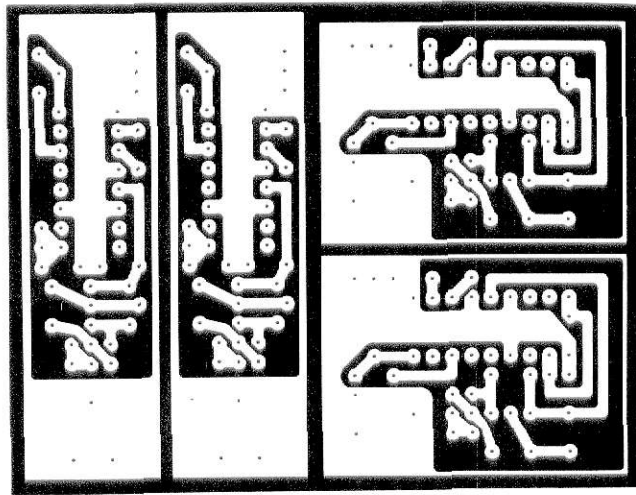


Fig. 4-1 Printed circuit conductor patterns for pulse modulated transmitter.

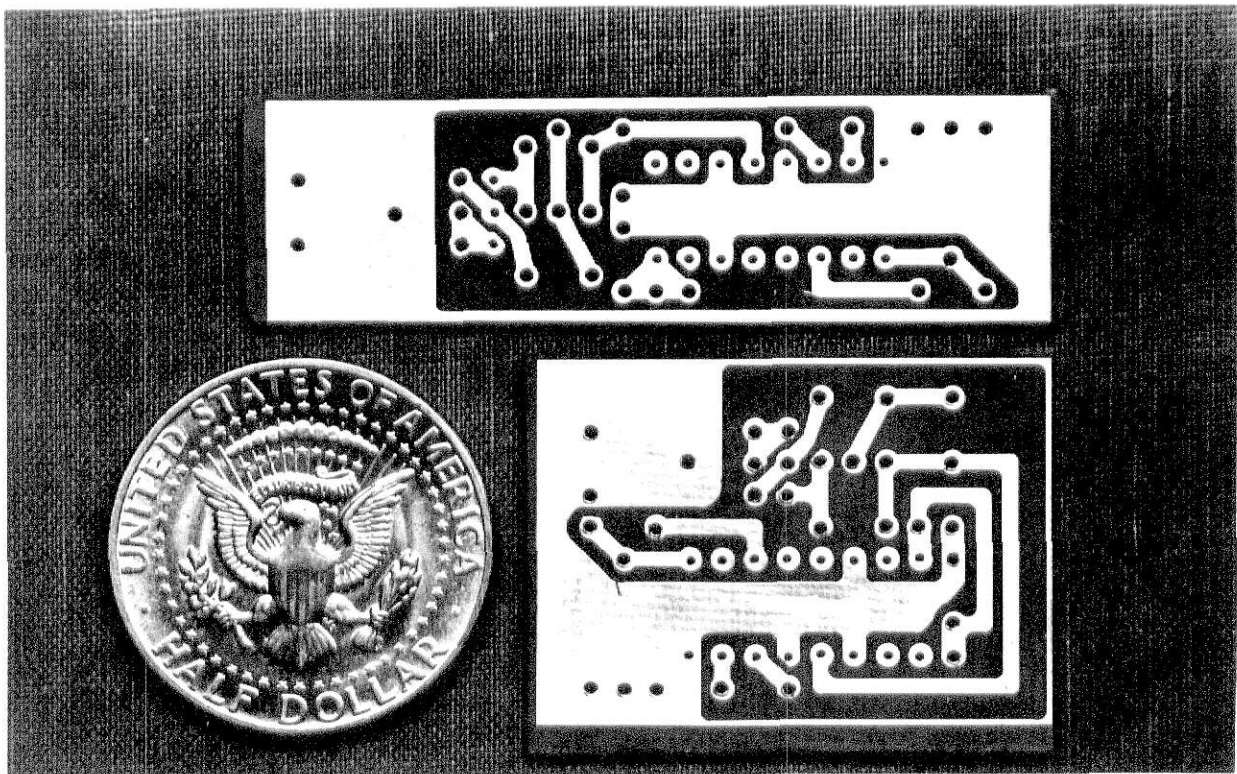


Fig. 4-2 Printed circuit boards for the pulse modulated transmitter.

The photoresist on the conductors was removed by rubbing with a paper towel soaked in methylene chloride. A light scrubbing with a steel wool pad insures the complete removal of any stubborn photoresist.

Next the conductors were tin plated. Tin plating the copper conductors facilitates good solder connections and prevents long term oxidation of the copper. The tin plating solution consists of the following:

Distilled Water	1 gallon
Stannous Chloride	2.5 ounces
Sodium Cyanide	25 ounces
Sodium Hydroxide	3 ounces

Time required for a 3" X 6" board to tin plate was approximately 10 minutes. The board was washed and dried; then the individual transmitter boards were separated by cutting them apart with a band saw. Eight transmitter boards were placed on each 3" X 6" board. The component lead holes were drilled with a 0.025" diameter bit for the transistor and integrated circuit leads and a 0.068" diameter bit for the other components. Fig. 4-2 shows examples of the finished boards. The components were then mounted, soldered, and tested to produce an operational telemetry transmitter.

One of the key questions in the evaluation of the telemetry transmitter design is "do we have an adequate signal from the

buried transmitter or is the signal so weak that it is undetectable." Field testing of a transmitter prototype yielded the answer. The test consisted of using a calibrated receiver to measure the received signal strength from the transmitter buried at various depths.

The first step was to calibrate a receiver with a known signal source. Due to fund limitations, a Sansui 800 Stereo FM receiver was used rather than a receiver of laboratory quality. However, the results are valid for making system performance calculations for high grade equipment. The Sansui 800 receiver sensitivity is rated by the manufacturer as 2 microvolts (-102 dbm for 50Ω load) for 20 db of quieting. The 20 db of quieting refers to the amount of noise reduction in the receiver output when an unmodulated signal of 2 microvolts is applied to the input. The receiver has a meter in the limiter circuit which is calibrated linearly from 0 to 5. The meter is normally used as a tuning aid, as it gives an indication of relative signal strength. It was found that 20 db of quieting was obtained with a signal strength of approximately 1 on the meter. The meter reading was calibrated in terms of signal voltage input at the antenna terminals with a Boonton Model 80 signal generator. The meter was found to be very nonlinear below about 0.5 and above 4.0. Therefore, only the range of 1 to 4 on the meter was calibrated. A step attenuater (Kay Electric Co. 432CB) was placed at the input of the receiver in the actual

tests to enable measurements of signals strong enough to saturate the receiver limiter stage. Fig. 4-3 shows the calibration curves obtained. The receiver was calibrated at 4 MHz intervals across the 88 to 108 MHz FM broadcast band as the frequency of an individual telemetry transmitter varies somewhat due to component tolerances. Fig. 4-4 shows the receiver sensitivity variation with frequency, which is anything but uniform.

The accuracy of this calibration technique suffered to a small degree by an impedance mismatch at the receiver input. The manufacturer claims that the receiver antenna input impedance is 75 ohms unbalanced. However, most RF test equipment, including the generator and step attenuator used in this project, are designed with 50 ohm impedance inputs and outputs. The 50 to 75 ohm impedance mismatch is not great in terms of the voltage standing wave ratio (VSWR) on the transmission line. It is 1.5 to 1. Additionally, the receiver input probably is not precisely 75 ohms resistive impedance and the means to measure it were not readily available. Therefore, it was decided to accept this deficiency rather than build an impedance transformation device to eliminate it.

A 100-foot length of RG-213 coaxial cable transmission line was employed to enable placing the receiver away from the location of the buried transmitter. This eliminated unwanted stray pick-up by the receiver and thus insured that the only signal entering the

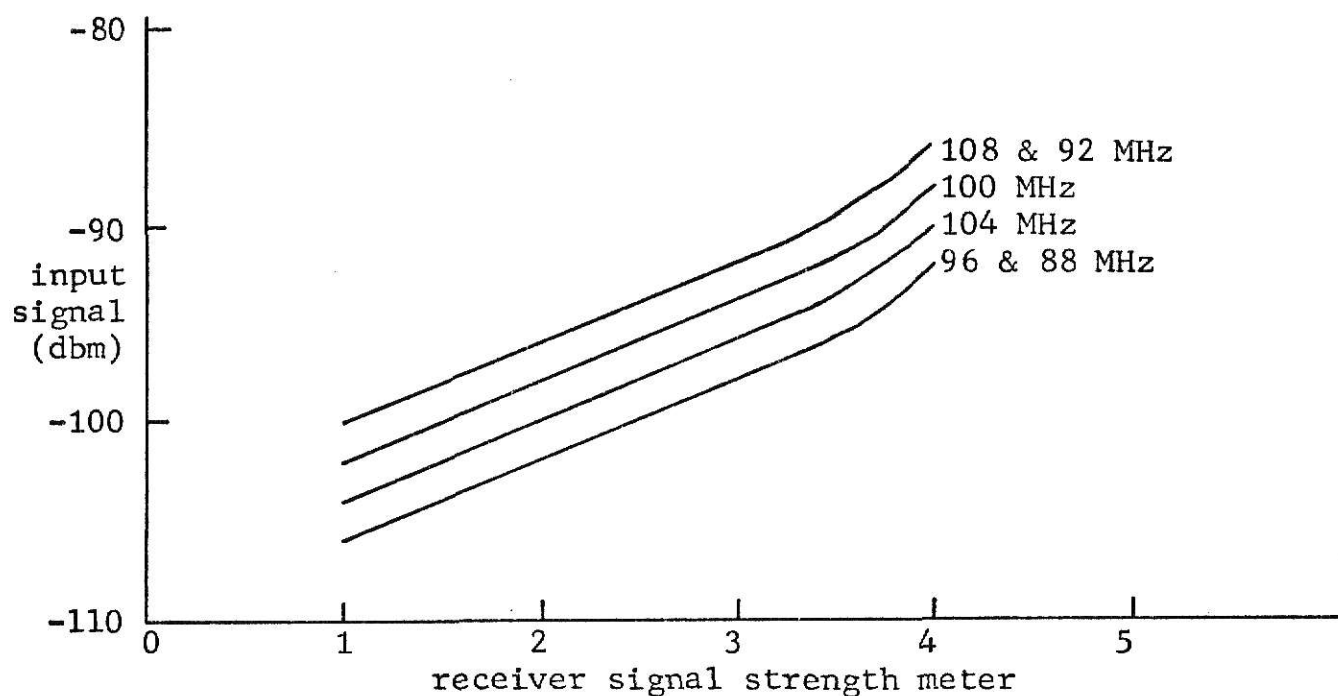


Fig. 4-3 Receiver calibration curves for the Sansui Model 800 FM receiver.

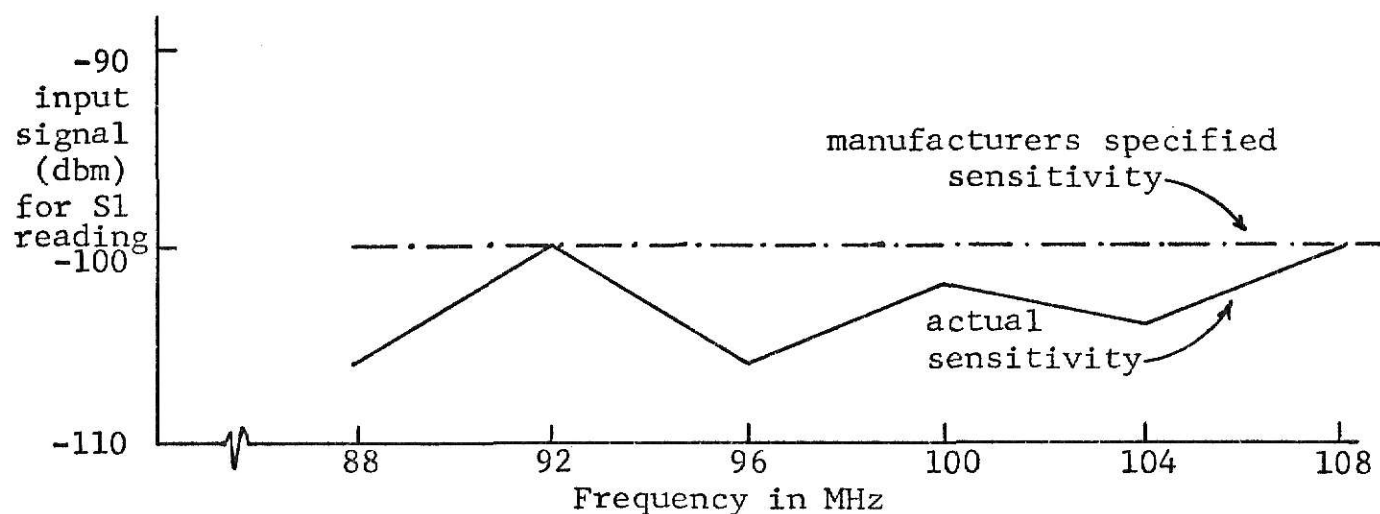


Fig. 4-4 Receiver sensitivity variation with frequency for Sansui Model 800 FM receiver.

receiver was via the dipole antenna and transmission line. The attenuation of 100 feet of RG-213 at 100 MHz is given as 2.2 db and the additional loss due to the impedance mismatch as 0.1 db [2]. Therefore, the total line loss is 2.3 db. During the calibration procedure, all readings were taken in dbm and were rounded off to the nearest whole decibel, thus 2 db was used as the line loss when computing the results.

The next step was to conduct the signal strength measurements from the buried transmitter. The transmitter without the modulator was constructed on one of the printed circuit boards. The transmitter was not modulated for these tests since the receiver was calibrated for an unmodulated signal. This allows the data to be applied to either the frequency modulated or pulse modulated telemetry transmitter with appropriate calculations.

The transmitter was enclosed in a small plastic box and attached to four "C" size carbon-zinc cells for the 6 volt power source. It was then buried in wet soil and the test set-up shown in Fig. 4-5 was employed to measure the signal strength. The curve on Fig. 4-6 shows the received signal strength as a function of the transmitter depth. The antenna height was one-foot above ground level. This height gave the highest readings for each burial depth of the transmitter. Fig. 4-7 shows a typical variation of signal strength versus antenna height for the case of the transmitter buried 14 inches in the ground. The antenna was

probably being detuned by the presence of the ground when the antenna was closer than one foot.

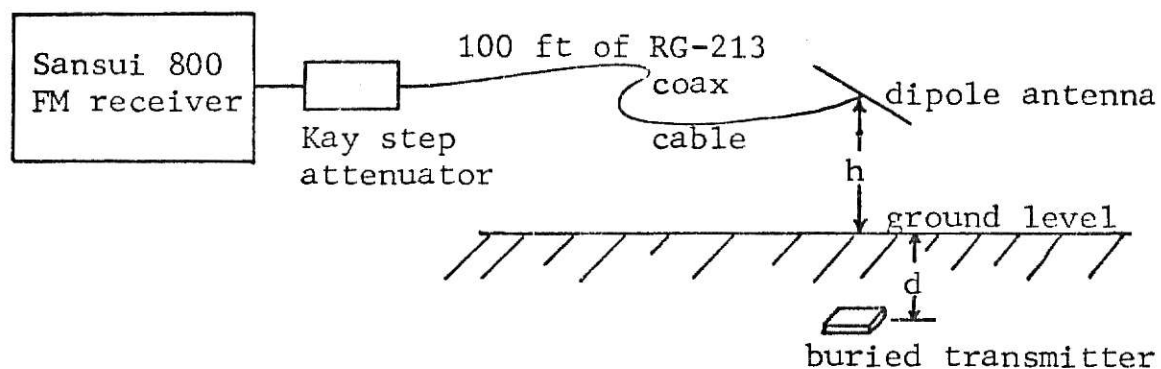


Fig. 4-5 Test setup for buried transmitter field strength measurements.

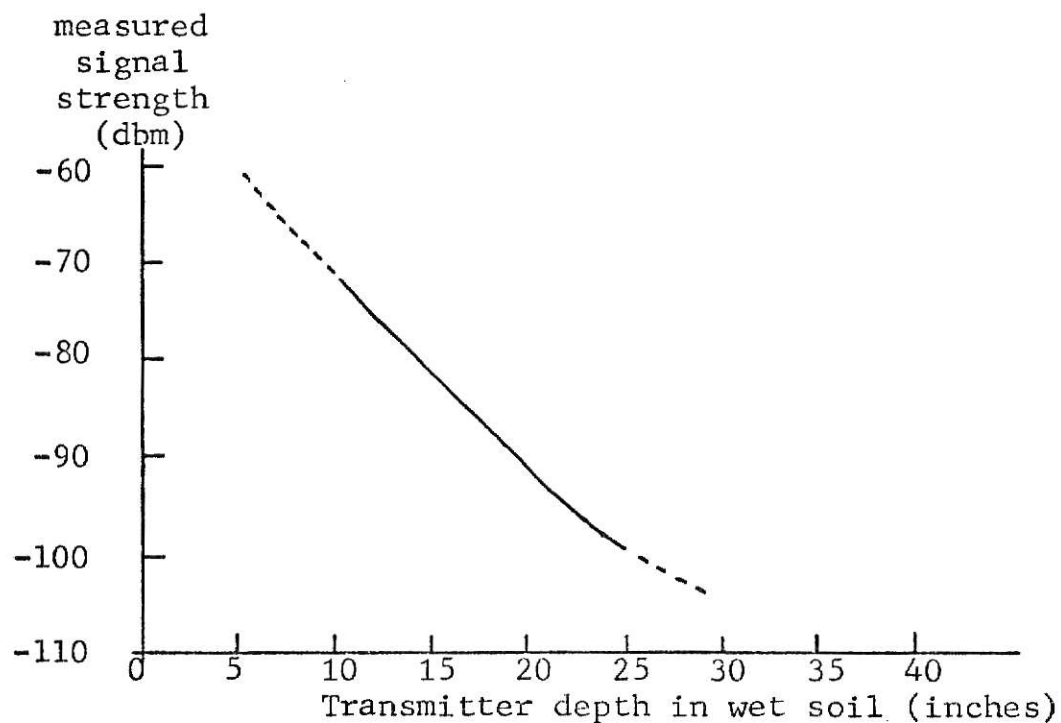


Fig. 4-6 Variation of received signal strength versus depth of transmitter burial.

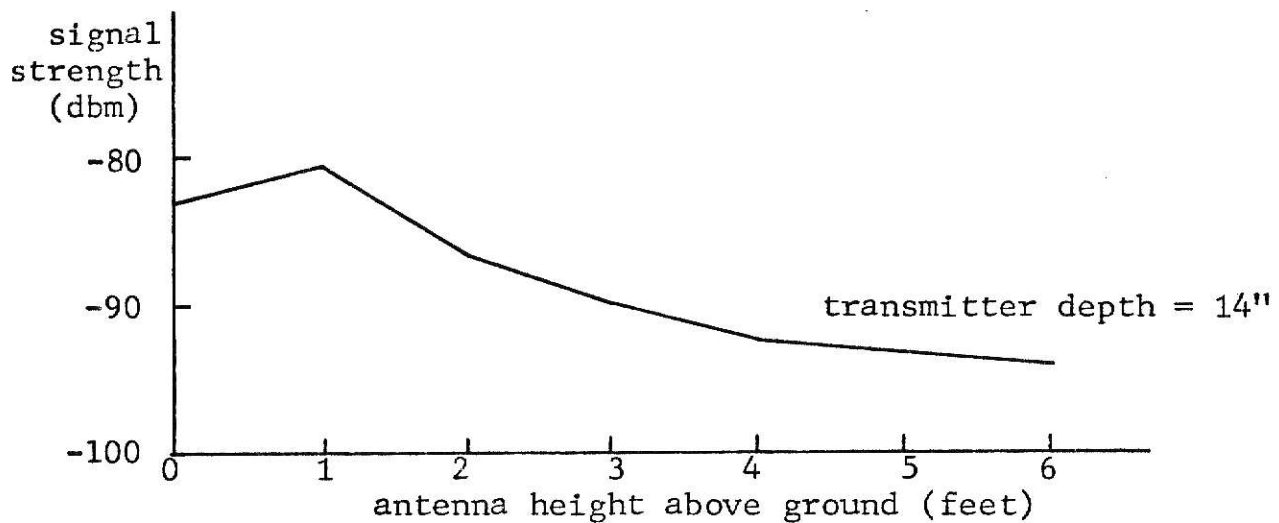


Fig. 4-7 Received signal strength versus antenna height.

A half-wave dipole antenna was used in these measurements. The length of the dipole was calculated from the formula $l \text{ (in.)} = \frac{5600}{f \text{ (MHz)}}$ [5]. A dipole 54 inches long (resonant at 104 MHz) was constructed and mounted on a 6-foot pole. The pole enabled the antenna to be easily positioned for the necessary measurements.

Another important question in evaluating the design of the telemetry transmitters is "what kind of sensor interface capability do the transmitters have and how well do they perform with a typical sensor." Since the design of the sensors or evaluation of all the available sensors coupled to the transmitter is beyond

the scope of this project, the information presented here will deal with the versatility of the transmitter in accommodating a wide variety of sensors and with results obtained using two sensors currently under development in the Department of Electrical Engineering at Kansas State University.

The control input to the CD4046AE VCO (Fig. 2-10) is the principal sensor input to the transmitter modulator. The control input (pin 9 of the CD4046AE) has an extremely high input resistance of approximately 10^{12} ohms. Thus it will have a negligible loading effect when connected to a high impedance sensor element. Since many sensor elements are a type of resistive element which varies with the parameter being measured, the sensor can be used in a simple high impedance voltage divider as shown in Fig. 4-8. The effective linear range of the VCO is from 10% to 90% of V_{DD} . The total resistance should not be higher than about 5 megohms to avoid noise pickup by the control input and not lower than about 500 K Ω to avoid excessive current drain from the batteries.

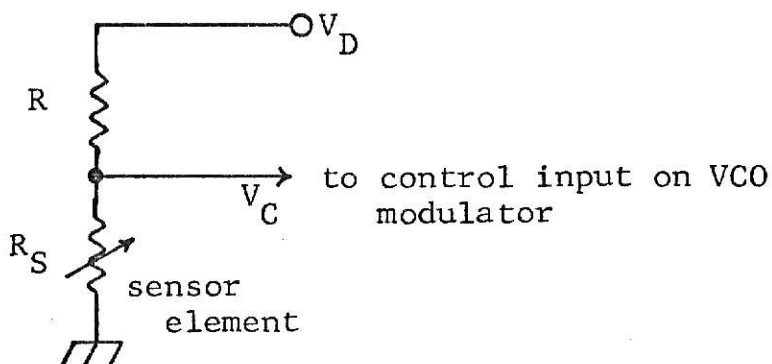


Fig. 4-8 Resistive sensor in voltage divider configuration.

We can apply these constraints to find the range of allowable sensor values.

$$V_C = \frac{R_S V_D}{R_T} \quad \text{where } R_T = R + R_S$$

$$\text{therefore: } R_T = R_S \frac{V_D}{V_C}$$

$$0.5 = R_S \left(\frac{V_D}{0.1V_D} \right)$$

$$R_S = 0.05 \text{ M}\Omega \quad \text{minimum } R_S$$

A similar calculation shows that the maximum value of R_S is 4.5 M Ω . Thus the modulator is capable of accepting quite a wide range of sensor resistance values when the voltage divider configuration is used. If a wider range is required, the sensor element must be placed in a bridge circuit. This will require a separate battery source be used to drive the bridge or a micro-power operational amplifier to interface the bridge circuit and the modulator input be used.

A KSU developed thick-film thermistor was connected to the telemetry transmitter to demonstrate temperature telemetry. The calibration curve of temperature versus modulator frequency for this particular unit is shown in Fig. 4-9. The data was taken by feeding the demodulated receiver output into a digital frequency counter.

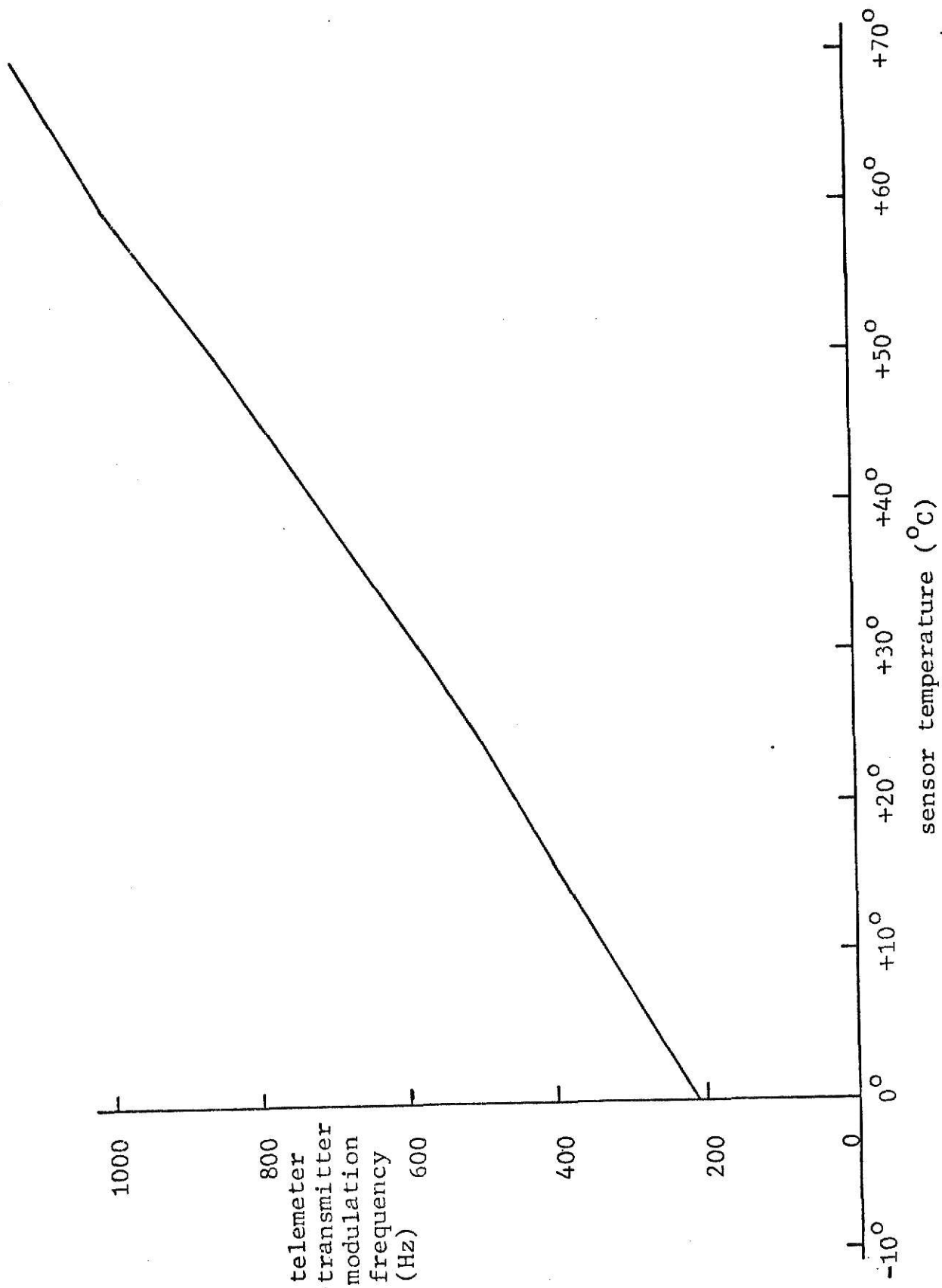


Fig. 4-9 Transmitter modulation frequency vs. temperature for KSU developed thermistor.

Another method of controlling the modulation frequency of the transmitter involves using a capacitive type sensor. In this case, the sensor is a nominal capacitance which varies with the parameter being measured. This type of sensor would be connected to pins 6 and 7 of the CD4046AE IC (Fig. 2-10), rather than the 200 pf capacitor. The control voltage input (pin 9) has to be connected to V_{DD} .

A KSU developed moisture sensor was interfaced with a telemetry transmitter and tested to demonstrate the telemetry capabilities of this type of sensor. The moisture sensor is actually a microcomb capacitor manufactured using thick-film techniques. It has a nominal capacitance of about 55 pf in air and 700 to 800 pf when immersed in water. These particular microcomb capacitors were constructed using conductor widths of 0.007-inch and a conductor spacing of 0.007 inches.

The microcomb capacitor sensor changes value and, hence, the modulation frequency whenever the dielectric constant of its surrounding medium changes. Since the dielectric constant of oven dried soil is about 2.6 compared to a dielectric constant for water of about 80, a significant change in capacitance should occur with a change in moisture content of the soil [9].

Measurements taken with the KSU developed capacitive moisture sensor are presented in Fig. 4-10. A number of measurements were taken at each soil moisture point and plotted. The curve represents the mean value of each set of measurements. The moisture content

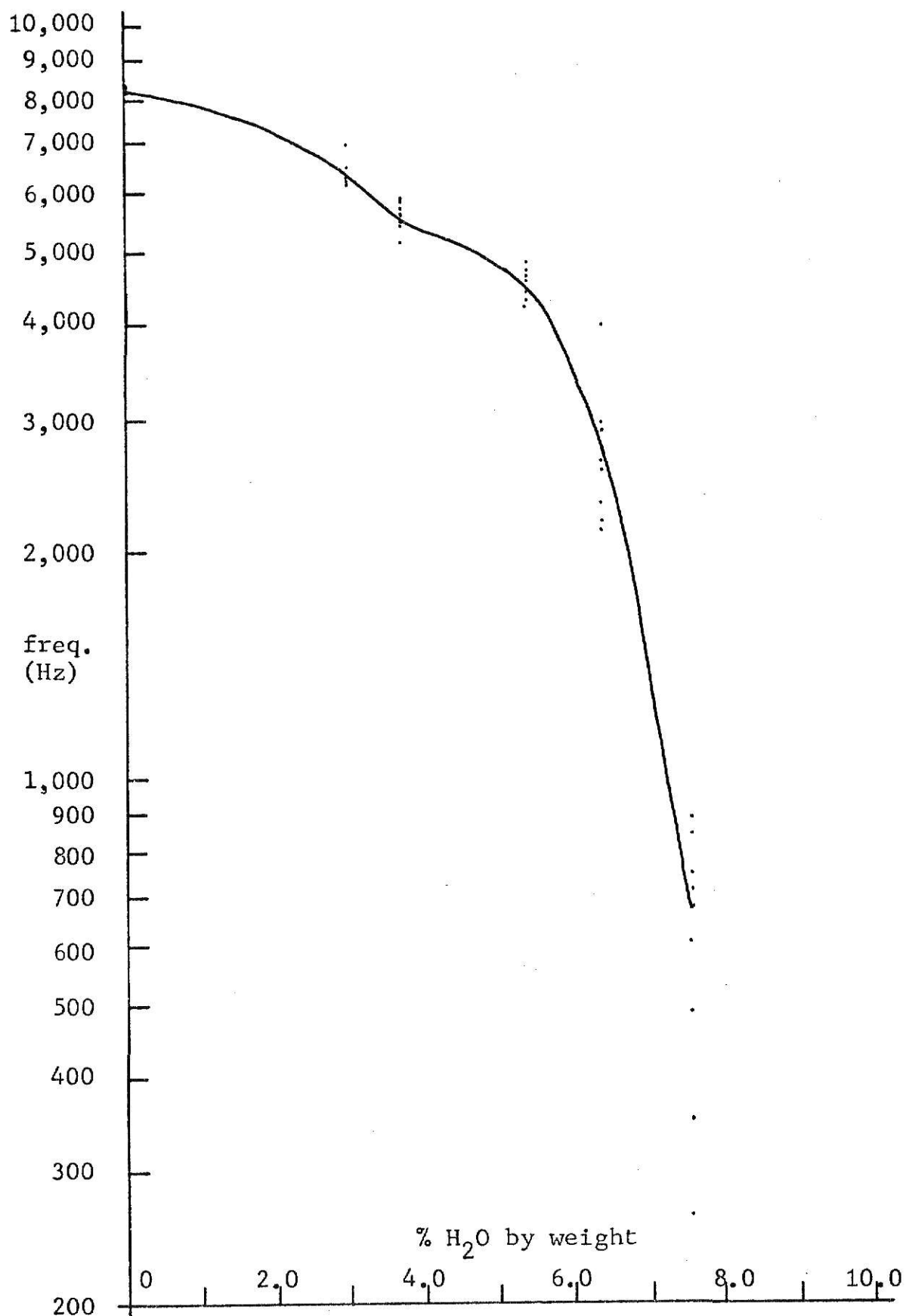


Fig. 4-10 Telemetry frequency vs. percent of soil moisture by weight.

of the soil was determined by the gravimetric method. The transmitter was implanted several times in a large container of soil. A reading was taken during each implantation. A sample of the soil was taken and weighed. The sample was dried at 160°C for 20 hours and reweighed. The percentage of soil moisture is the difference in weights divided by the weight of the wet soil sample.

The sensor was very sensitive to soil moisture as can be seen from the curve of Fig. 4-10. The large variation of readings taken for a particular sample were probably due to differences in the soil sensor interface. This effect became quite noticeable at the higher moisture contents.

CHAPTER V

CONCLUSIONS

The pulse modulated transmitter, rather than the frequency modulated transmitter, is probably the best solution to the highway roadbed telemetry problem. It's major advantages are lower power consumption, fewer components, no adjustable components, and lower cost. However, the frequency modulated telemetry transmitter can be used with readily available home entertainment FM receivers. The pulse modulated signal can be demodulated on such a receiver; but, as pointed out in Chapter II, optimum results can be obtained by the use of a special receiver with a narrower bandwidth and an AM detector. The frequency modulated transmitter is shown in Fig. 5-1 and the pulse modulated version is shown in Fig. 5-2.

The extremely low power consumption (85 to 90 μ a) of the pulse modulated transmitter makes very long transmitter lifetimes possible with currently available batteries. An accurate prediction of the lifetime using the Lithium size "D" cells is difficult as the 8-ampere hour capacity gives a lifetime (neglecting for the moment the 25% increase and 10% decrease in capacity factors mentioned in Chapter II) of 94,118 hours or 10.74 years. The manufacturer's rated shelf life is given as 10 years. Thus we can only predict that the transmitter life should approach the shelf life of the batteries. The transmitter life for the currently available sizes

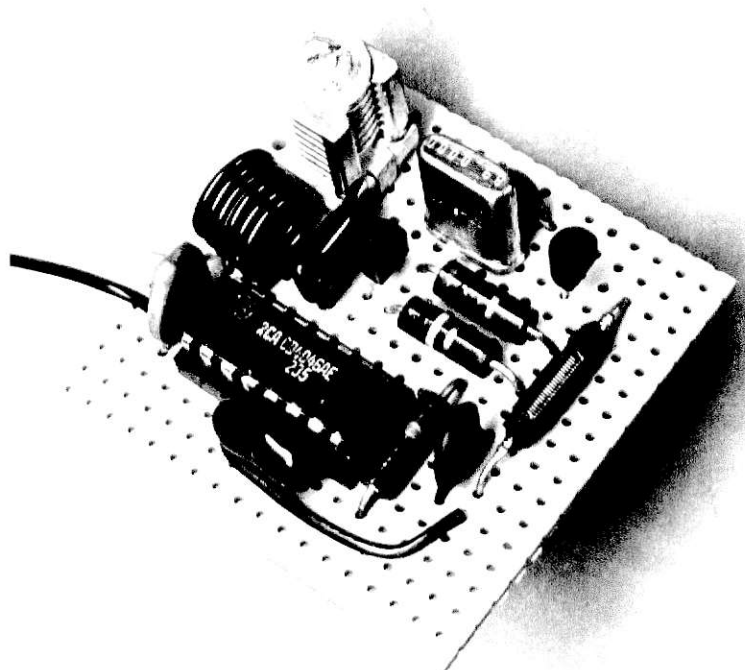


Fig. 5-1 Frequency modulated telemetry transmitter.



Fig. 5-2 Pulse modulated telemetry transmitter with moisture sensor and Lithium batteries.

of Lithium batteries are tabulated as follows:

<u>Cell Size</u>	<u>Cost for two cells</u>	<u>Rated Capacity</u>	<u>Transmitter Lifetime</u>
D	\$22.96	8.0 ampere-hour	10.74 years
C	\$15.88	3.2 ampere-hour	4.30 years
3/4C	\$11.80	2.5 ampere-hour	3.36 years
2XRM1	\$ 5.00	1.0 ampere-hour	1.34 years

Thus, for applications requiring a shorter lifetime, the cost and physical size of the telemetry package may be reduced. For example, the 2XRM1 cell is only 1.325 inches long and 0.635 inches in diameter.

The fact that the pulse modulated transmitter uses fewer components is incidental to the design. The fewer components used (consistent with other design requirements) results in the smallest physical size of transmitters. One printed circuit layout required an area of only 1.7 square inches. Additionally, the probability of a transmitter malfunction due to component failure has to be lower with fewer components. The two components which would lower the reliability of the frequency modulated transmitter and are not required on the pulse modulated version are the crystal and the variable capacitor in the tuned drain circuit. The crystal is a relatively fragile component, even when it is manufactured to military specifications. The variable capacitor, which is required to peak the oscillator output on the crystal's frequency, is also

a problem when designing a physically rugged circuit. It would present special problems when encapsulating the transmitter. If the capacitor was totally encapsulated or potted, future adjustments would be impossible. One could not be sure that the adjustment would remain peaked and at such a low power level, anything less than optimum could spell disaster when the transmitter was buried. If the capacitor adjustment were to be made available after encapsulation, the problem of adequate encapsulation and potting would become magnified.

The small quantity costs of each type of telemetry transmitter, excluding sensors, is tabulated below.

<u>Item</u>	<u>Pulse Modulated Transmitter</u>	<u>Frequency Modulated Transmitter</u>
CD4046AE IC	\$ 6.50	\$ 6.50
MPF102 Transistor	.90	.90
R2501 Epicap Diode	----	1.10
90 MHz Crystal	----	8.15
Two "D" Size Lithium Cells	22.98	22.98
Misc. parts	<u>3.00</u>	<u>3.50</u>
TOTAL	\$33.38	\$43.13

The major portion of each transmitter's cost is the presently expensive Lithium cells. This cost can be reduced drastically if one elects to forego the low temperature capability and long battery life of the Lithium cell and use a common carbon-zinc or

alkaline cell. The case for the Lithium cell is stronger when one considers that the cost per transmitter per year would be approximately \$3.50 (pulse modulated version). Eventually the Lithium cells and the COSMOS IC will be available for much less. Thus, a transmitter cost of under \$10.00 in small quantities should be realizable in the near future.

The other crucial factor in the success of the telemetry transmitter design is the RF output power. The transmitter must radiate sufficient energy from the circuit wiring to be received with an adequate signal-to-noise ratio when the transmitter is buried up to 24 inches in the ground. Tests show that these transmitters meet this requirement. Assuming a sensitive receiver with a sensitivity of $0.1 \mu\text{v}$ for 10 db signal-to-noise ratio were used, the transmitters could probably be buried up to 36 inches or more.

A modification of the pulse modulated transmitter to wideband frequency modulation could probably be successful. The current consumption would rise to about 140 microamperes, but the signal could be effectively received on a standard FM broadcast receiver. The modulation would probably have to be classified as wideband FM since with this type of oscillator and modulator circuit, it would be difficult to achieve a small enough deviation for narrow band FM. If narrow band FM were achieved, the receiver would have to remain relatively wideband as the oscillator stability would not be good enough (without reverting back to crystal control) for

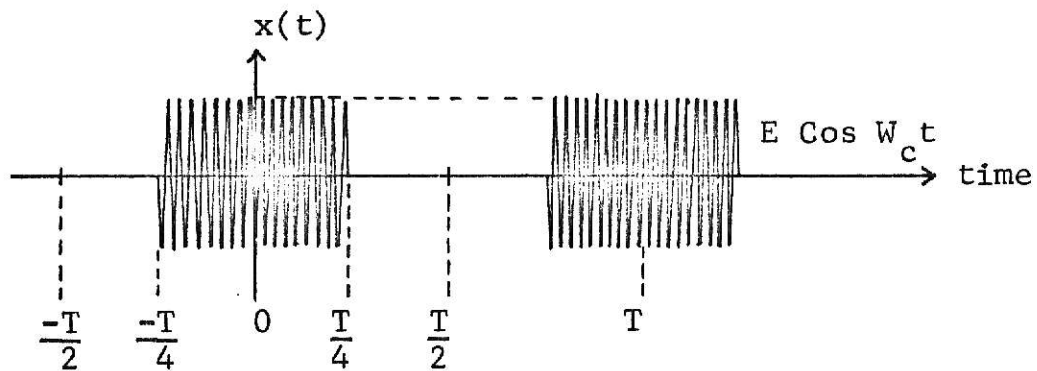
narrow band reception. Other modifications or variations on the design presented may become apparent and, for certain specific sensor applications, may be desirable. However, this design should be close to optimum for a very wide range of applications and sensors. A proper receiver designed with the necessary bandwidth, detector, and data display would complete the telemetry system.

When specific highway measurement programs are identified, the proper sensors can be designed, fabricated, interfaced with the transmitter, and calibrated. The complete telemetry system can then be deployed for actual operation.

APPENDIX

Calculation of Required Receiver Bandwidth for Pulse Modulation

The transmitter is a pulse modulated RF carrier which may be represented in the time domain as:



where the carrier is $E \cos W_c t$ and the modulation square wave is symmetrical with period T .

$$\text{Thus: } x(t) = \frac{1}{T} \sum_{k=-\infty}^{\infty} C_k e^{jW_k t} ;$$

$$\text{where } W_k = \frac{k2\pi}{T} , \quad k = 0, 1, 2, \dots$$

in the Fourier series expansion.

The spectrum of the modulated signal can be found by evaluating the C_k terms where:

$$C_k = \int_T x(t) e^{-jW_k t} dt = \int_{-T/4}^{T/4} E \cos W_c t e^{-jW_k t} dt$$

Substituting: $\cos W_c t = \frac{e^{jW_c t} + e^{-jW_c t}}{2}$

$$C_k = \int_{-T/4}^{T/4} E \left(\frac{e^{jW_c t} + e^{-jW_c t}}{2} \right) e^{-jW_k t} dt$$

$$C_k = \frac{1}{2} \left[\int_{-T/4}^{T/4} E e^{-j(W_k - W_c)t} dt + \int_{-T/4}^{T/4} E e^{-j(W_k + W_c)t} dt \right]$$

$$C_k = \frac{1}{2} \left[\frac{E}{-j(W_k - W_c)} \left(e^{-j(W_k - W_c) T/4} - e^{+j(W_k - W_c) T/4} \right) \right.$$

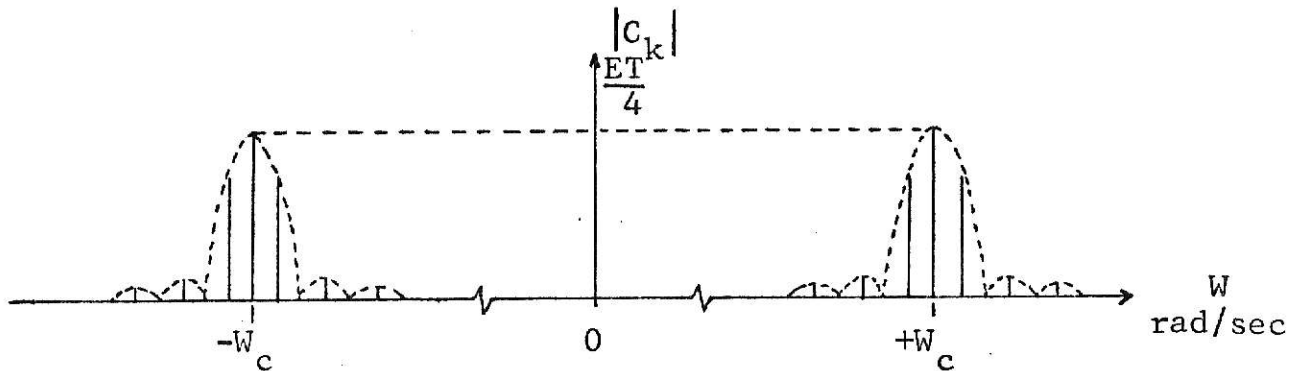
$$\left. + \frac{E}{-j(W_k + W_c)} \left(e^{-j(W_k + W_c) T/4} - e^{+j(W_k + W_c) T/4} \right) \right]$$

Rearranging terms

$$C_k = \frac{1}{2} \left[\frac{ET \left(e^{j(W_k - W_c) T/4} - e^{-j(W_k - W_c) T/4} \right)}{4j(W_k - W_c) T/4} + \frac{ET \left(e^{j(W_k + W_c) T/4} - e^{-j(W_k + W_c) T/4} \right)}{4j(W_k + W_c) T/4} \right]$$

$$C_k = \frac{1}{2} \left[\frac{ET}{2} \cdot \frac{\sin(W_k - W_c) T/4}{(W_k - W_c) T/4} + \frac{ET}{2} \cdot \frac{\sin(W_k + W_c) T/4}{(W_k + W_c) T/4} \right]$$

The spectrum of the signal can be seen by plotting $|C_k|$ versus frequency



Note: spectrum lines are spaced $\frac{2}{T}$ rad/sec.

Most of the power is concentrated in the carrier and first pair of sidebands. Therefore, if the receiver bandwidth were $\frac{4\pi}{T}$ rad/sec wide at the -3 db points, the receiver will recover the modulation with close to optimum signal-to-noise ratio. Thus, the

receiver bandwidth should be twice the highest modulation frequency of the transmitter.

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A TELEMETRY TRANSMITTER
FOR HIGHWAY DATA COLLECTION

by

RODERICK KIRK BLOCKSOME

B. S., Kansas State University, 1968

AN ABSTRACT OF A MASTER'S THESIS

submitted in partial fulfillment of the

requirements for the degree

MASTER OF SCIENCE

Department of Electrical Engineering

KANSAS STATE UNIVERSITY
Manhattan, Kansas

1973

A brief outline of the general problems of short range telemetry from a highway roadbed is presented. Information from similar previous work is examined and its applicability to the telemetry of highway sensors is discussed. The major steps in the evolution of a satisfactory telemetry transmitter design are presented. Particular emphasis is placed on the relationship of transmitter power consumption, radio frequency (RF) output power, and transmitter lifetime in arriving at the final design.

A temperature sensor and a moisture sensor, both fabricated with thick film techniques, are discussed. The versatility of the thick film thermistor as a temperature sensor is pointed out. The microcomb capacitor moisture sensor is discussed and a brief explanation of its operation is presented.

Two types of telemetry transmitters were designed; one uses frequency modulation and the other uses pulse modulation. The relative merits of each are discussed and receiver requirements for each are pointed out.

Experimental data are presented on received signal strength from a buried transmitter and the testing procedures for these measurements are described. Sensor interface with the telemetry transmitter is discussed and experimental data from a temperature sensor and a moisture sensor are presented.