

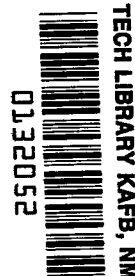
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## A SPACECRAFT ANALOG-TO-DIGITAL CONVERSION SYSTEM EMPLOYING MOSFETS

*by Donald C. Lokerson*

*Goddard Space Flight Center  
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NATIONAL AERONAUTICS AND SPACE ADMINISTRATION

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## ABSTRACT

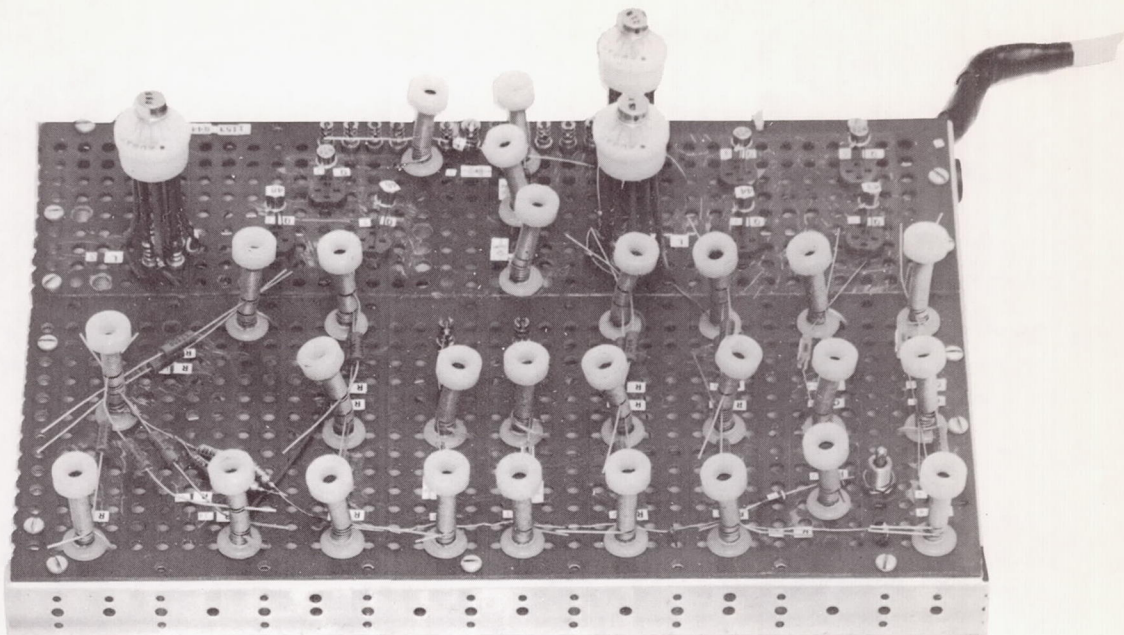
The analog-to-digital conversion systems used in past Interplanetary Monitoring Platform spacecraft are reviewed to indicate why a new voltage-controlled oscillator is required and what characteristics are desirable for the new oscillator. Since the system developed employs Metal Oxide Semiconductor Field Effect Transistors (MOSFETS), this type of transistor is described.

Analysis, design, and developmental innovations are described in detail for the system, which is composed of an analog commutator, a voltage-to-frequency oscillator, an internal voltage regulator, and a digital storage register. The system performance using special calibration equipment is described in detail for the operation over a nominal 0 to +5 volt input with a 5-millisecond sample time. The linearity is within  $\pm 0.5$  percent of the 6-kHz to 46-kHz frequency range. Tests indicate long-term stability is about  $\pm 0.2$  percent over a three-month period, whereas short-term stability is  $\pm 0.5$  percent after warmup.

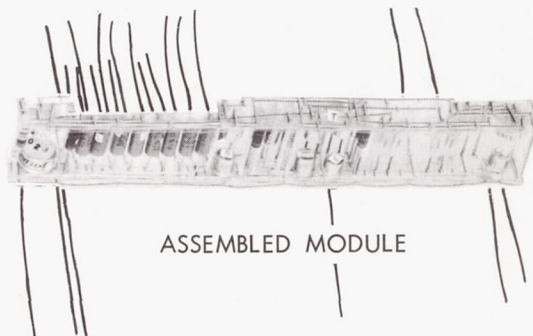
The system described can be adapted to permit measurements at hundreds of megohms impedance. The time constants can be changed to allow a wide range of frequencies and sample times. The applications for this analog-to-digital conversion appear quite broad and may find use where microminiaturization and moderate sample times are desired.

## CONTENTS

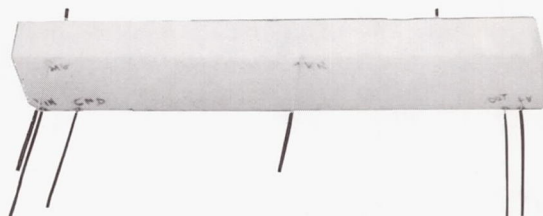
Abstract . . . . .	ii
INTRODUCTION. . . . .	1
HISTORY OF THE DEVELOPMENT OF THIS ANALOG-TO-DIGITAL CONVERTER. .	1
SYSTEM REQUIREMENTS. . . . .	2
THE MOSFET DEVICE. . . . .	5
THE ANALOG COMMUTATOR. . . . .	7
THE ANALOG OSCILLATOR. . . . .	7
THE VOLTAGE REGULATOR . . . . .	12
THE DIGITAL ACCUMULATOR. . . . .	15
THE ANALOG OSCILLATOR SYSTEM PERFORMANCE . . . . .	16
Test Equipment . . . . .	18
Final Analog Oscillator Circuit Calibration. . . . .	20
Long-Term Stability . . . . .	21
Power Dissipation . . . . .	24
Mechanical Layout . . . . .	24
SUMMARY OF RESULTS AND CONCLUSIONS. . . . .	25
ACKNOWLEDGMENTS . . . . .	26
Bibliography . . . . .	26



CALIBRATION BREADBOARD



ASSEMBLED MODULE



POTTED MODULE

Frontispiece—The analog oscillator in three stages of assembly.

# **A SPACECRAFT ANALOG-TO-DIGITAL CONVERSION SYSTEM EMPLOYING MOSFETS**

by  
Donald C. Lokerson  
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## **INTRODUCTION**

The Interplanetary Monitoring Platform (IMP) spacecraft contain sensors which produce voltages proportional to temperature, current, and other parameters; these voltages must be commutated into an analog-to-digital converter. This conversion is performed by gating the output of a voltage-controlled oscillator into a binary accumulator for a fixed time period. Because this conversion averages the sensor voltage over the sample period, greater immunity to noise can be achieved than with the more conventional successive-approximation analog-to-digital converter.

Recent advances in the production of Metal Oxide Semiconductor Field Effect Transistors (MOSFETS) have allowed the IMP analog-to-digital conversion system to take advantage of the excellent properties of these devices. They have extremely high-input-impedance gates and are easily integrated on one chip in large quantities. The analog commutator, voltage-controlled oscillator, and accumulator in the system are mainly composed of integrated circuit MOSFETS.

The system to be described in detail commutates analog voltages from ground to +5 volts into a linear voltage-to-frequency oscillator which can be gated "on" for 5.0 milliseconds. The output frequency of 6 kHz to 46 kHz is accumulated by an 8-bit binary counter which is completely integrated on one chip.

A history of similar analog-to-digital converters will be provided as a background for comparison of such earlier systems with the new system. Then a detailed discussion of the new system, to be used in future IMP spacecraft, will serve as one example of the possible applications of MOSFETS in analog-to-digital conversion.

## **HISTORY OF THE DEVELOPMENT OF THIS ANALOG-TO-DIGITAL CONVERTER**

Previous IMP spacecraft employed pulse-frequency modulation (PFM) channel coding of both analog and digital data. Voltage-controlled oscillators were employed to produce these pulse frequencies directly from the analog voltages. A 5-kHz to 15-kHz magnetic-core multivibrator was selected for long-term stability. On IMP D and E, an 8-bit accumulator was used to digitize some of the analog data by sampling the oscillator frequency for 20 milliseconds.

Future spacecraft will require analog-to-digital conversions as frequently as every 5 milliseconds, thus a new oscillator will be required. The difficulties in temperature compensation of the magnetic-core oscillator and the desire to cover a much wider frequency range increase the desirability of other methods of controlling the oscillator frequency. The IMP D and E oscillator required presetting an accumulator and allowing it to overflow once before the normal readout

occurred. Thus it could "underflow" if the frequency became too low. Avoiding such a possibility was one objective of the new design.

The first MOSFET analog oscillator studied employed a constant current discharge of a variable-height sawtooth; however, short-term stability and linearity were poor because of difficulty in controlling the sawtooth resetting.

The second approach employed the MOSFET circuit shown in Figure 1. As  $V_{IN}$  is made more negative, the effective impedance of  $L_3$  decreases, thus increasing the oscillator frequency. The MOSFET changed impedance too quickly to operate over the 5-volt range desired; thus further development was not considered.

The third MOSFET analog oscillator proved the most promising, because linearity and temperature coefficients could be well compensated over a very wide range of frequency. This oscillator will be described in detail.

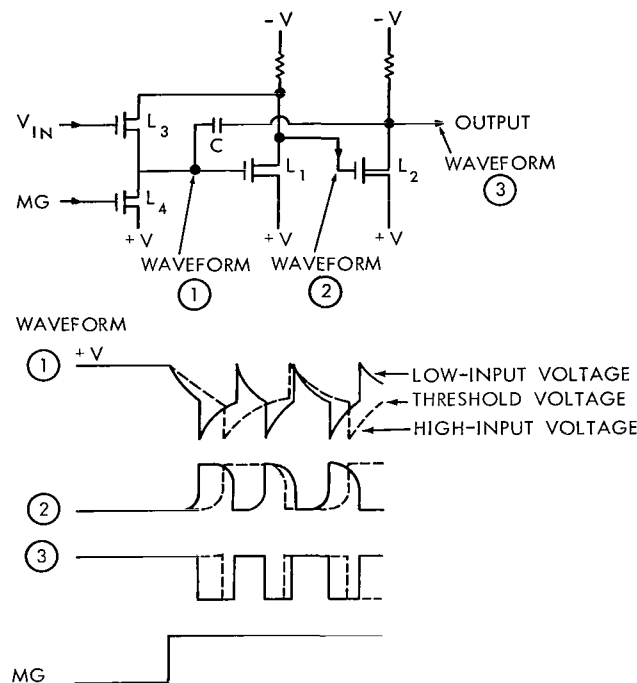


Figure 1—Second MOSFET oscillator considered.

A class of linear-period oscillators was discounted because the distortion in transfer function is detrimental, as shown in Figure 2. The digitization accuracy is not constant across the voltage range, a result which is undesirable. Figure 2 puts a temperature sensor voltage response through the analog-to-digital conversion both for the linear-period and for the linear-frequency transformation, with the dashed lines showing the distortion produced by the linear-period oscillator.

## SYSTEM REQUIREMENTS

The design goals for the IMP H, I, and J spacecraft analog-to-digital conversion are:

1. Normal input voltage range = 0 to +5 volts
2. Operating input voltage range = -0.5 to +5.5 volts
3. Digitization resolution = 0.5%
4. Calibration and linearity over normal voltage range:

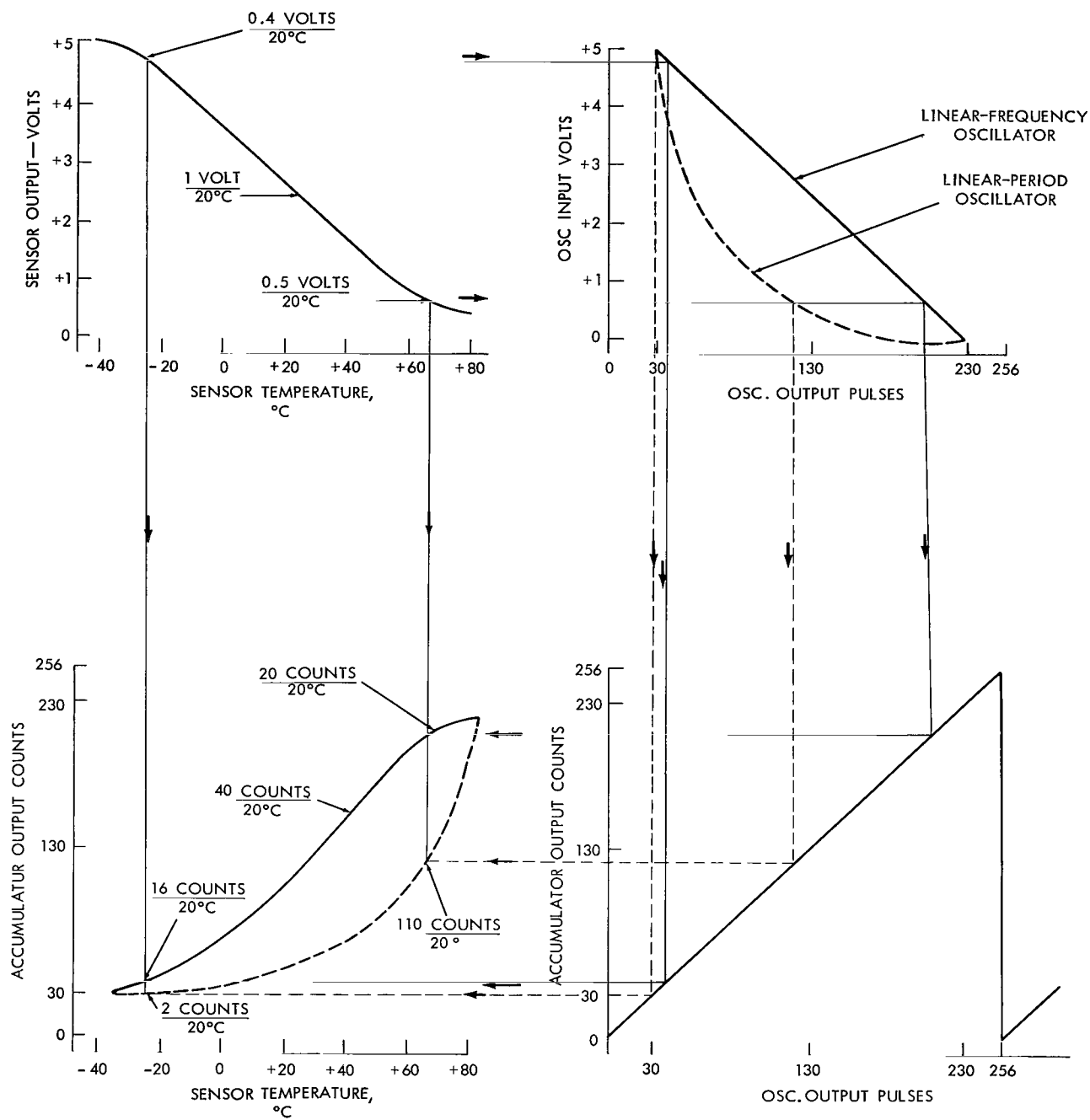


Figure 2—Effect of oscillator linearity.

- a. +10°C to +30°C =  $\pm 1/2$  count
- b. -5°C to +45°C =  $\pm 1$  count
- c. -35°C to +85°C =  $\pm 3$  counts



5. Voltage input loading by analog oscillator:
  - a. less than .04 microamperes at ground
  - b. more than 5 megohms at 5 volts
6. Ground-to-5-volt square wave input voltage response to be within  $\pm 2$  counts of 2.5-volt input calibration
7. First-day-after-turn-on stability =  $\pm 2$  counts
8. Short-term stability =  $\pm 1/2$  count
9. Long-term stability =  $\pm 2$  counts
10. Consistency between oscillators over the  $-5^{\circ}$  to  $+45^{\circ}\text{C}$  range =  $\pm 1$  count
11. Analog sample period = 2.5 ms or 5.0 ms

An 8-bit accumulator has a digital output range of 256 counts, or slightly better than 0.5-percent resolution. The oscillator nominal voltage input range of 0 to +5 volts requires an output range of 200 counts, but a "safety margin" of 0.5 volts would be desirable. The 56 extra counts should be divided nearly equally above and below the nominal extremes, thus defining the frequency of the oscillator as

$$\frac{30 \text{ counts at 5 volts}}{2.5 \times 10^{-3} \text{ seconds per sample}} = 12 \text{ kHz at 5 volts}$$

and

$$\frac{230 \text{ counts at 0 volts}}{2.5 \times 10^{-3} \text{ seconds per sample}} = 92 \text{ kHz at 0 volts.}$$

A similar calculation for a 5-millisecond sample time yields a 6- to 46-kHz range for the oscillator.

The oscillator should operate nearly linearly over a large range, from +5.5 volts to -0.5 volts, so that, since

$$\frac{200 \text{ counts range}}{5 \text{ volts range}} = \frac{20 \text{ counts}}{0.5 \text{ volts "safety margin"}}$$

it follows that

$$\frac{10 \text{ counts at 5.5 volts}}{2.5 \times 10^{-3} \text{ seconds per sample}} = 4 \text{ kHz at +5.5 volts}$$

and

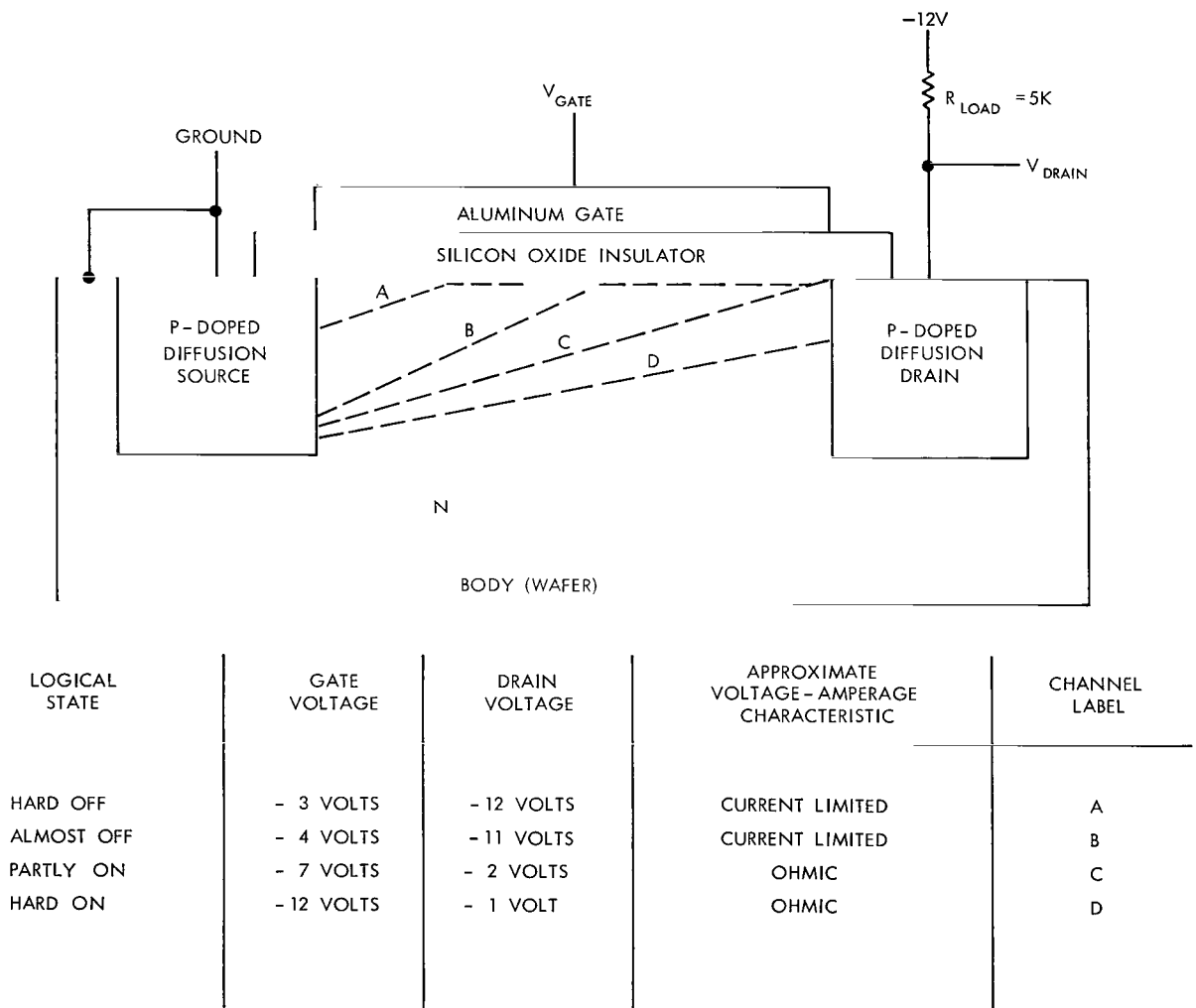
$$\frac{250 \text{ counts at -0.5 volts}}{2.5 \times 10^{-3} \text{ seconds per sample}} = 100 \text{ kHz at -0.5 volts.}$$

A similar calculation for a 5-millisecond sample time yields a 2-to-50-kHz range for the oscillator. This voltage-controlled oscillator must operate approximately linearly over nearly 5 octaves!

The input commutator may have up to 50 inputs commutated into the one analog oscillator. The oscillator output may be gated into several binary accumulators for data storage until readout by the spacecraft telemetry.

# THE MOSFET DEVICE

A detailed description of the Metal Oxide Semiconductor Field Effect Transistor (MOSFET) is necessary because of its particular properties, and because many similar devices are produced by industry. The "P" channel MOSFET is constructed on an N-type body with the P-doped diffusions, as shown in Figure 3. A layer of silicon dioxide insulates an aluminum "gate" from the device, producing an input impedance of hundreds of megohms. This property is utilized for the commutator and oscillator input gates. As the gate is made more negative than the more positive P region, a channel is formed, as shown in Figure 3, producing characteristics as shown in Figure 4, when the MOSFET is used as an inverter. The MOSFET is used as a saturated inverter switch in the analog commutator and in the digital circuits in the accumulator.



(ASSUME 4-VOLT THRESHOLD DEVICE)

Figure 3—The MOSFET as an inverter.

Part of the analog oscillator utilizes the MOSFET as an emitter follower, or source follower. Figure 5 shows the change in gate-to-source voltage as a function of current through the device. In general any linear change in the oscillator with voltage input can be calibrated out, but any non-linearity is difficult to cancel out. The non-linear change in this device is about 10 percent over the entire range of interest (10 to 250 microamps), but is less than 1 percent if the two piece-wise approximations can be made. As explained later, the analog oscillator will achieve this result by employing two different source followers, both determining the frequency when less than +3 volts is put into the oscillator. Only one source follower determines the frequency when more than about +3 volts is put into the oscillator. Slightly different

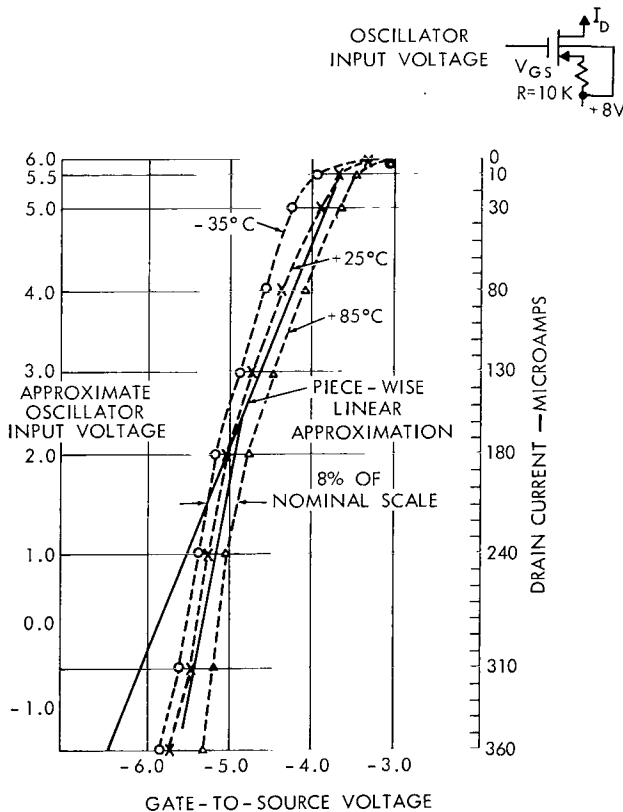


Figure 5—Source follower MOSFET characteristic.

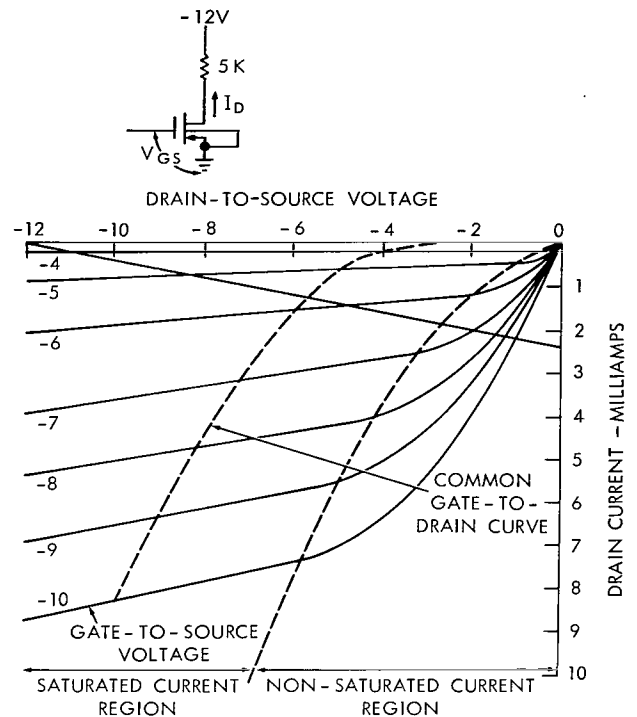


Figure 4—Inverter MOSFET drain characteristic at room temperature.

values for  $R$  in the two source followers will produce a nearly linear result.

The temperature effects on the MOSFET when used in the source follower circuit are quite large, but nearly linear, as shown in Figure 5. The 0.15 volt decrease in gate-to-source voltage from  $-35^{\circ}\text{C}$  to  $+25^{\circ}\text{C}$  is about equal to the 0.15 volt decrease from  $25^{\circ}\text{C}$  to  $85^{\circ}\text{C}$ . This drop in voltage will be compensated for by a corresponding voltage drop in the +8 volt regulator supply as the temperature increases. When the gate and drain are maintained at nearly equal voltages, the MOSFET gate-to-source voltage varies linearly with temperature, as shown in Figure 6. This property will be utilized in the MOSFET regulator as described later.

The MOSFETS used in the analog oscillator almost completely determine the voltage-to-frequency characteristics of the oscillator, and

it is only because of the particular voltage-to-current characteristics of the MOSFETS that the analog oscillator described later can be made linear and temperature-compensated to a high degree.

## THE ANALOG COMMUTATOR

The IMP spacecraft encoding systems will include an analog data processor, which multiplexes the outputs of voltage monitors, temperature monitors, and experiment sensors into one analog-to-digital converter, as shown in Figure 7. Three performance parameters, PP1, PP2, and PP3, are shown in this simplified example. The resistors  $R_{IN}$  are 100 kilohms to 500 kilohms, and represent a constant load impedance to the experiment, with the remainder of the commutator appearing as an insignificant additional load. The resistor  $R_{LIMIT}$  is about 1 kilohm to limit the current drawn if a very high positive voltage is put into the terminal PP1.

Diode  $D_1$  clamps the voltage so that MOSFET  $L_1$  will be able to turn off by application of +8 volts to the gate of  $L_1$ . This prevents cross-talk to other sensors being commutated. The body lead of  $L_1$  must be more positive than the +8.7 volts to which the source is clamped, to prevent leakage. When commutation of PP1 into the analog-to-digital converter is desired, square waves A, B, and C are all negative, turning that branch of the logic matrix "on." The positive voltage on the gate of  $L_4$  turns device  $L_4$  "off." The -10 volts applied to  $L_1$  gate is sufficient to turn the  $L_1$  device hard "on" even if the input PP1 is at ground. The source-to-drain impedance of the  $L_1$  device is about 1 kilohm when the gate is more than 6 volts more negative than the source. The source-to-drain impedance of the  $L_1$  device is hundreds of megohms when the gate is +8 volts.

As the waveforms in the logic matrix change, several microseconds are required for the new analog voltage to be gated "on," because of the high resistor values, which conserve power. The analog-to-digital converter will not begin taking a sample until at least 2.5 milliseconds after the new analog data line switching has begun. Thus this circuit conserves power by allowing non-critical switching times. Negligible additional loading (hundreds of megohms) of the sensor develops during the commutation of the analog voltage.

## THE ANALOG OSCILLATOR

The MOSFET source follower characteristics produce a 4-percent non-linearity current output due to the non-linear change in gate-to-source voltage with current. The current as shown in Figure 8 will determine the frequency of a multivibrator, producing approximately the same non-linearity

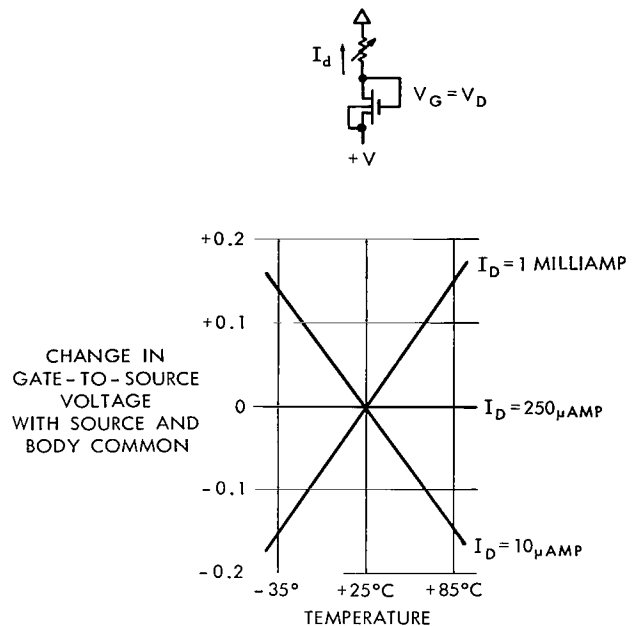


Figure 6—Ideal variation in gate-to-source voltage with temperature.

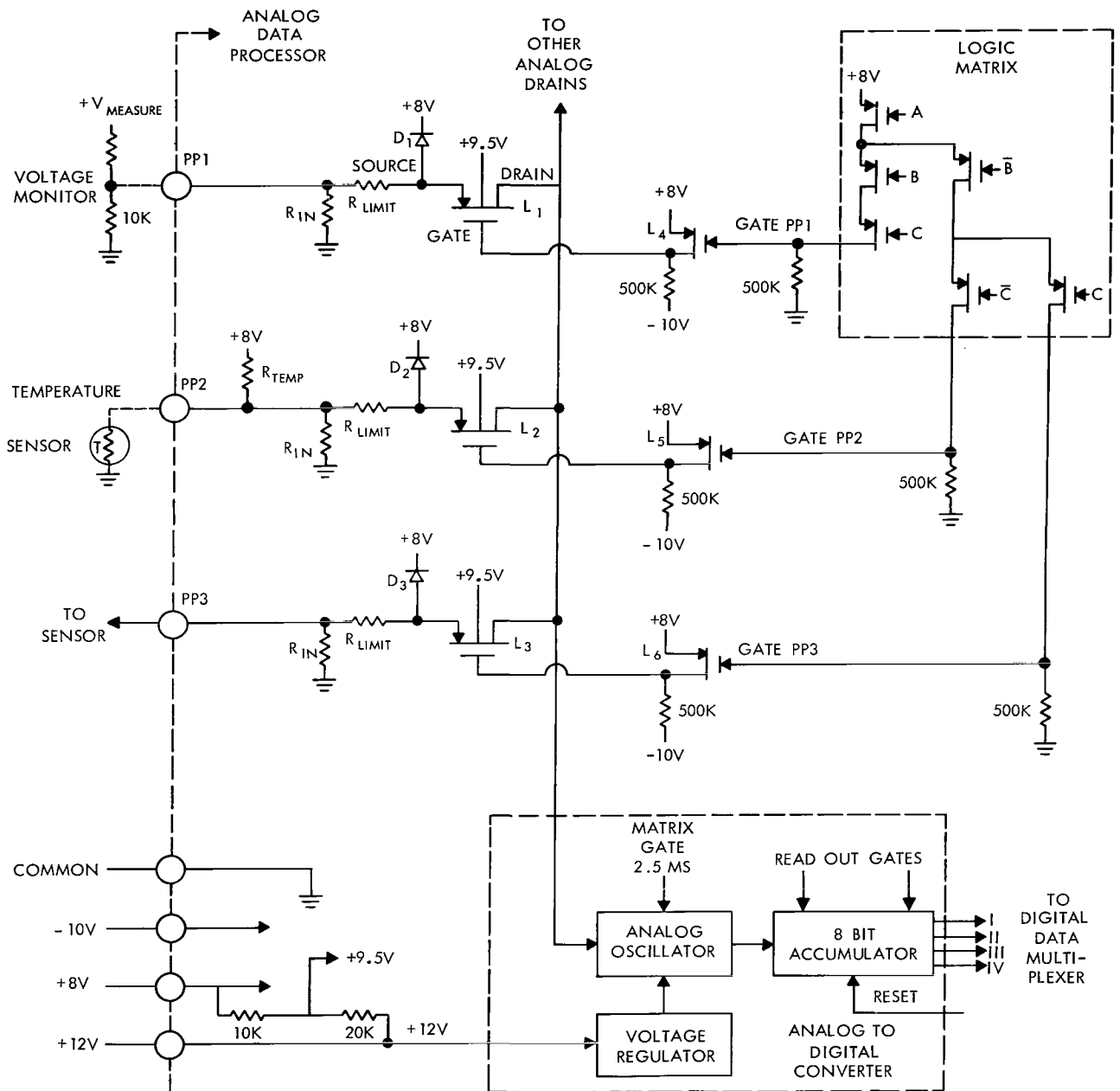


Figure 7—Simplified analog data processing system.

in frequency. The operation of the multivibrator begins with  $Q_1$  "off" and  $Q_2$  "on." The current  $I_1$  will charge  $C_1$  in some time period  $\tau_1$  until  $Q_1$  is turned "on." This forces the base of  $Q_2$  to go negative almost +V volts (-8.82 volts, in this instance). This turns  $Q_2$  "off," causing the base of  $Q_1$  to go even more positive, thus reinforcing the switching action. Then  $I_2$  charges  $C_2$  during a time period  $\tau_2$ , until  $Q_2$  is turned "on" again, thus turning  $Q_1$  "off." This switching action returns the oscillator to the original state. Since the input voltage-to-current characteristic is non-linear, a similar non-linearity

will occur in the voltage-to-frequency characteristic, producing a higher than desired frequency at high currents, as shown in Figure 9. However, if  $R_1$  is made infinite, and  $R_3$  provides a nearly constant current to  $C_1$ , the output frequency will approach  $1/\tau_1$  as  $I_2$  increases and  $\tau_2$  approaches zero. Note that this is the opposite effect from when  $I_1$  supplied current; thus a combination of  $I_1$ ,  $I_2$ , and  $I_3$  can be selected to approach a linear characteristic. This is shown mathematically as approximately:

$$\frac{i_n}{C_n} = \frac{dv_{\text{saw tooth}}}{dt}$$

or

$$\int_0^{\tau_n} dt = \int_{v_{sw_n}}^{+0.5V} \frac{C_n}{I_n} dv,$$

$$T_n = \frac{C_n}{I_n} (v_{sw_n} + 0.5),$$

$$f_{\text{OUT}} = \frac{1}{\tau_1 + \tau_2 + \tau_{\text{SWITCHING}}} = \frac{1}{C_1(v_{sw} + 0.5) \frac{1}{I_1 + I_3} + C_2(v_{sw} + 0.5) \frac{1}{I_2}},$$

where in practice

$$|v_{sw_n} + 0.5| \approx |V|,$$

$$C_1 \approx C_2 \approx C,$$

$$V_1 \approx V_2 \approx V_{\text{IN}} + V_{\text{GS}},$$

( $V_{\text{IN}}$  = oscillator input volts,

$V_{\text{GS}}$  = gate-to-source voltage as determined by the source current)

$$T_{\text{SWITCHING}} \rightarrow 0,$$

$$I_1 = \frac{V - V_{\text{IN}} - V_{\text{GS}}}{R_1},$$

$$I_2 = \frac{V - V_{\text{IN}} - V_{\text{GS}}}{R_2},$$

$$I_3 \approx \frac{3V}{2R_3} \text{ (averaged),}$$

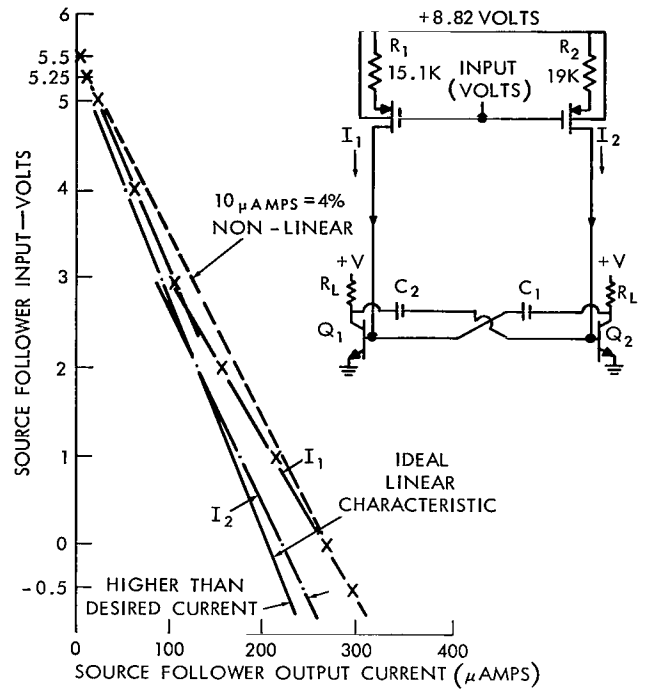


Figure 8—MOSFET source follower input to analog oscillator.

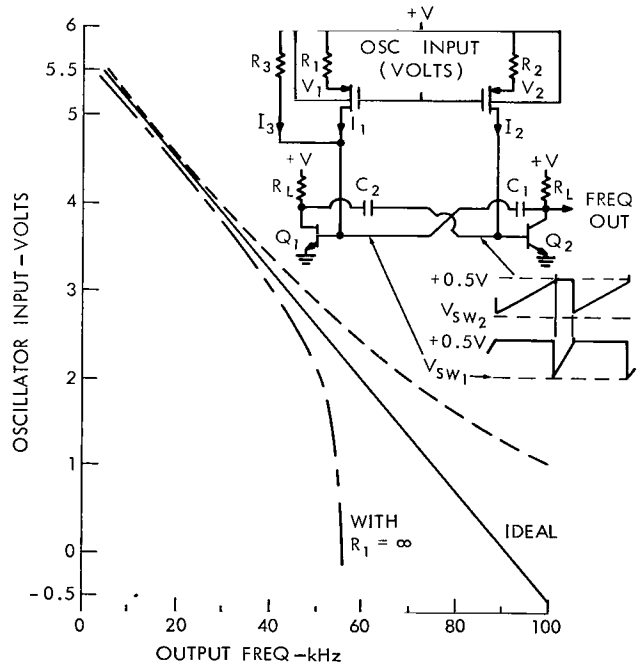


Figure 9—Linearity correction characteristic.

and

$$f_{OUT} \approx \frac{1}{V \left( \frac{C}{I_1 + I_3} + \frac{C}{I_2} \right)} = \frac{1}{CV \left( \frac{1}{I_1 + I_3} + \frac{1}{I_2} \right)} = \frac{V - V_{IN} - V_{GS}}{CV \left( \frac{1}{R_1} + \underbrace{\frac{3V}{2R_3(V - V_{IN} - V_{GS})}}_{\text{LINEARITY CORRECTION TERM}} + R_2 \right)} ;$$

or, if  $I_3$  is negligible,

$$f_{OUT} \approx \frac{1}{CV \left( \frac{1}{I_1} + \frac{1}{I_2} \right)} = \frac{1}{CV \left( \frac{R_1 + R_2}{V - V_{IN} - V_{GS}} \right)} = \frac{V - V_{IN} - V_{GS}}{CV(R_1 + R_2)} .$$

Thus, the non-linear input MOSFETS voltage-to-current characteristic is canceled by the non-linear effect to the multivibrator of the linearity correction term  $I_3$ .

As the input voltage,  $V_{IN}$ , increases, the frequency output decreases. Any change in threshold voltage,  $V_{GS}$ , produces the same amount of change in frequency as an equal change in the input voltage,  $V_{IN}$ , would produce. The same change in the supply voltage,  $V$ , will change the output frequency only about 2/3 as much because the sawtooth amplitude changes to help compensate. Since  $V_{GS}$  will change because of temperature, radiation, and other effects, the supply voltage,  $V$ , will make a corresponding change, such that

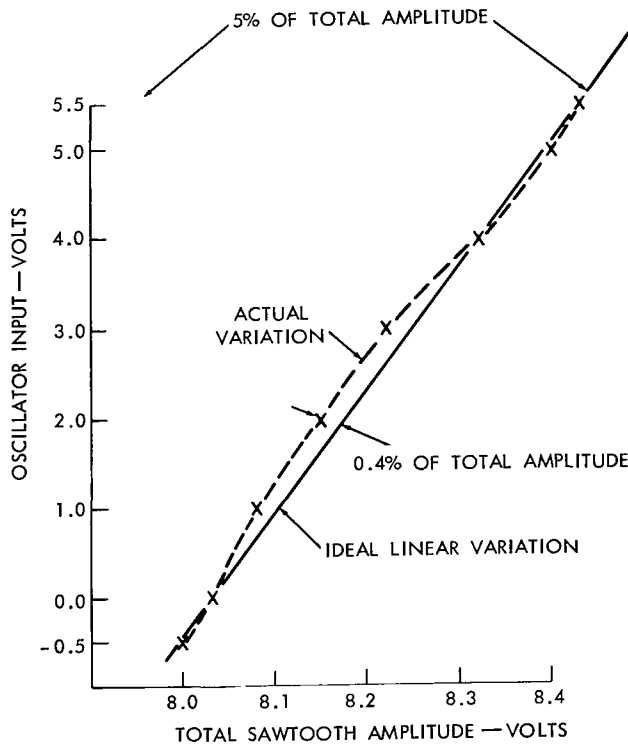


Figure 10—Multivibrator sawtooth variations.

$$f_{OUT} \approx \frac{(V + \Delta V_{GS}) - V_{IN} - (V_{GS} + \Delta V_{GS})}{K} \approx \frac{V - V_{IN} - V_{GS}}{K} .$$

A MOSFET within the voltage regulator will perform this function.

The multivibrator is not a perfect current-to-frequency converter; thus some of these characteristics must be considered. As the currents  $I_1$  and  $I_3$  increase, the transistors  $Q_1$  and  $Q_2$  will be held on harder, increasing the storage time at the moment of switching. The total sawtooth voltage amplitude,  $V_{sw}$ , varies because of this effect, as shown in Figure 10. A small increase in the  $Q_1$  and  $Q_2$  base diode drops also occurs with increased current. The total sawtooth amplitude plotted in Figure 10 is nearly linear; thus this variation is calibrated out of the oscillator.

When low frequencies are generated at high input voltages, the current required to turn  $Q_1$

or  $Q_2$  "on" is dependent on the value of the load resistors  $R_L$ . Since the oscillator should still operate with less than 5 microamperes flowing into the  $Q_2$  base at 5.5 volts, a complementary multivibrator was substituted for the simpler version, as shown in Figure 11. The 0.1 microampere steady state base current into  $Q_7$  and  $Q_8$  is capacitively bypassed to provide very fast multivibrator switching at very low frequencies. If a voltage higher than +6.0 volts should exist for even a short instant, the multivibrator would stop operating, because of lack of current  $I_2$ . The ability to resume multivibrating when the signal drops again to within normal limits is doubtful, and it depends on the fall time of the signal when within the normal voltage range again. To prevent such a problem, a minimum multivibrator frequency is determined by resistor  $R_4$ , in Figure 11.

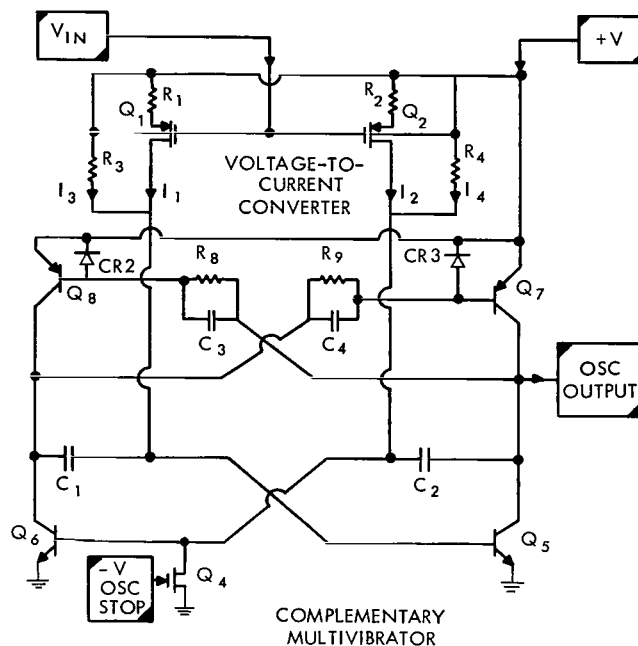


Figure 11—Complete oscillator circuit.

The operation of the entire oscillator can now be understood. The analog-to-digital conversion of the voltage  $V_{IN}$  is accomplished by gating the oscillator output frequency into an accumulator. To achieve the maximum resolution, of plus 1 count and minus no counts, the initial phase of the oscillator operation should be at the same point at the beginning of the sample. If MOSFET  $Q_4$  in Figure 11 is held "on" for 2.5 milliseconds, then  $Q_6$  will be held "off" while  $I_3$  and  $I_1$  will hold  $Q_5$  "on." This will turn  $Q_8$  "on" and  $Q_7$  "off." At the time a sample is to begin,  $Q_4$  is quickly turned "off" by pulling "-V osc stop" above ground. Then current  $I_2$  and  $I_4$  will charge capacitor  $C_2$  until  $Q_6$  begins to turn "on." This will turn  $Q_7$  "on" and  $Q_5$  "off," thus reinforcing the  $Q_6$  "turn-on" process, until the switching is complete, with  $Q_5$  and  $Q_8$  "off" and  $Q_7$  and  $Q_6$  "on." The base of  $Q_5$  is now at  $-(+V)$  and slowly increases as  $C_1$  is charged by  $I_1$  and  $I_3$ , until  $Q_5$  begins to turn "on" and the circuit returns to the original state, except that  $Q_6$  base is then at  $-(+V)$ . Oscillation continues at a frequency determined by  $I_1$ ,  $I_2$ ,  $I_3$ , and  $I_4$  until stopped by a negative voltage at "-V osc stop."

The capacitors  $C_3$  and  $C_4$  are large enough to fully switch  $Q_8$  and  $Q_7$ , with only  $R_8$  and  $R_9$  holding the circuit in the steady state. CR2 and CR3 clamp the turn-off transient of  $Q_8$  and  $Q_7$  respectively, so that only 1 volt base change is required to turn the  $Q_8$  or  $Q_7$  "on."

The "-V osc stop" voltage must be about -9 volts to turn  $Q_4$  "on" well. If  $V_{IN}$  is a large negative voltage,  $I_2$  becomes quite large, and the internal "on" resistance of  $Q_4$  could allow  $Q_6$  to turn "on" when not desired. Thus a limit to the negative voltage possible is desirable. The MOSFET commutator will prevent more than a -5 volt input to the oscillator.

The characteristics of the multivibrator over a temperature range of  $-35^{\circ}\text{C}$  to  $+85^{\circ}\text{C}$  are mainly dependent on the timing capacitors  $C_1$  and  $C_2$ , since the transistor gain changes do not affect the



saturated logic of the complementary switch multivibrator significantly. Consequently  $C_1$  and  $C_2$  are very stable, high-quality capacitors. The linear temperature characteristic of these capacitors, as shown in Figure 12, has only a 0.1-percent non-linearity in capacitance change with temperature. The +140 parts per million per degree centigrade temperature coefficient will be calibrated out by the oscillator power regulator. The resistors employed in the voltage-to-current generator change less than 25 parts per million per degree centigrade from  $-55^{\circ}\text{C}$  to  $+175^{\circ}\text{C}$ . The 0.3-percent change in resistance is small enough to neglect over the normal operating temperature range.

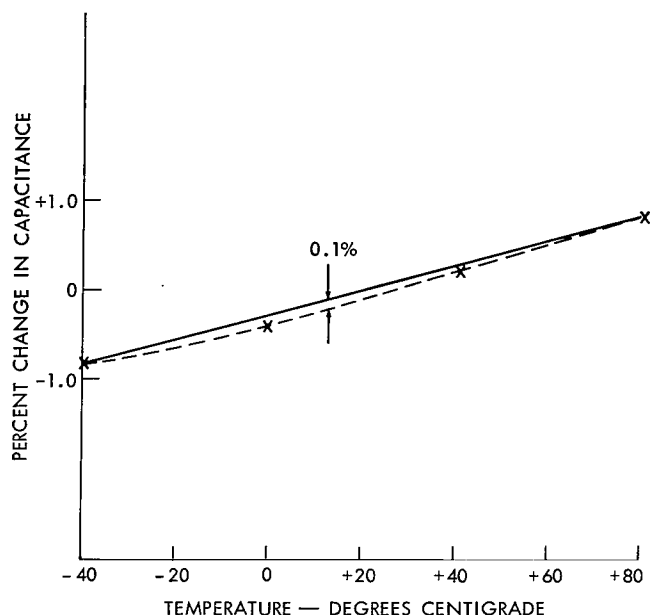


Figure 12—Capacitor temperature characteristic.

## THE VOLTAGE REGULATOR

The analog oscillator frequency depends greatly on the supply voltage from which it is powered. This voltage is adjusted to determine the frequency at +5.0 volts, and must change with temperature to compensate for the temperature coefficients of the MOSFETS and capacitors. Any small change in threshold voltage in the input MOSFETS must be compensated for by the power regulator. Thus a voltage reference using a MOSFET was chosen, with the original circuit shown in Figure 13a. The input voltage  $V_P$  is around 12 volts, and the required regulated voltage is about 8 volts; thus  $Q_{11}$  drops the voltage difference of about 4 volts. The regulated voltage is divided by  $R_5$  and  $R_6$  to produce a voltage  $V_1$ . The MOSFET  $Q_3$  acts like a poor zener diode voltage source when its gate and drain are connected together, as was shown in Figure 4; thus  $V_2 = V_{\text{REG}} - V_{\text{GS}}$ . The difference amplifier  $Q_9$  and  $Q_{10}$  amplifies any difference between  $V_1$  and  $V_2$ , and increase or decrease the current to correct the voltage difference. If the oscillator input voltage decreases, the current increases, and the regulated voltage decreases slightly by some  $\Delta V$ . The original voltages are indicated as unprimed symbols, and voltages after the change in current are primed nomenclature, in the equations below.

Assuming that  $Q_9$  and  $Q_{10}$  loading can be neglected:

$$V_1 = V_{\text{REG}} \left( \frac{R_6}{R_5 + R_6} \right),$$

$$V_2 = V_{\text{REG}} - V_{\text{MOS}},$$

where  $V_{MOS}$  is the source-to-drain voltage;

$$V_1' = (V_{REG} + \Delta V) \left( \frac{R_6}{R_5 + R_6} \right),$$

$$V_2' = (V_{REG} + \Delta V) - V_{MOS}',$$

and

$$\begin{aligned} \Delta(V_1 - V_2) &\approx (V_1 - V_1') - (V_2 - V_2') = \left( \frac{R_6}{R_5 + R_6} \right) (V_{REG} - V_{REG} - \Delta V) \\ &\quad - (V_{REG} - V_{MOS} - V_{REG} - \Delta V + V_{MOS}') \\ &\approx - \left( \frac{R_6}{R_5 + R_6} \right) (\Delta V) + \Delta V + V_{MOS} - V_{MOS}'. \end{aligned}$$

Where  $V_{MOS} \approx V_{MOS}'$  and where  $R_5 \approx R_6$ ,

$$\Delta(V_1 - V_2) \approx +\Delta V \left( 1 - \frac{R_6}{R_5 + R_6} \right) \approx +\Delta V \left( \frac{1}{2} \right).$$

Thus in this circuit  $Q_3$  is used as a zener diode, without utilizing the amplification available in the device. The regulation was improved by using the circuit shown in Figure 13b. The MOSFET  $Q_3$  now amplifies the change in gate-to-source voltage, which is

$$\Delta V_{GS} = \left[ (V_1' - V_1) - (V_{REG} - V_1) \right] = (V_{REG} + \Delta V) \left( 1 - \frac{R_6}{R_5 + R_6} \right) - V_{REG} \left( 1 - \frac{R_6}{R_5 + R_6} \right) = +\Delta V \left( 1 - \frac{R_6}{R_5 + R_6} \right),$$

so that the change in voltage at the difference amplifier, as previously determined, is still

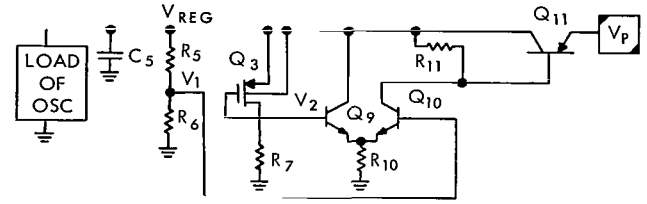
$$\Delta(V_1 - V_2) \approx - \left( \frac{R_6}{R_5 + R_6} \right) (\Delta V) + \Delta V + V_{MOS} - V_{MOS}';$$

but now

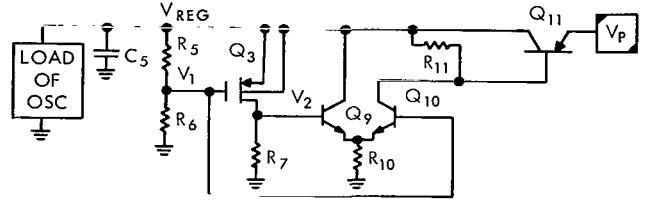
$$V_{MOS}' \neq V_{MOS},$$

but instead

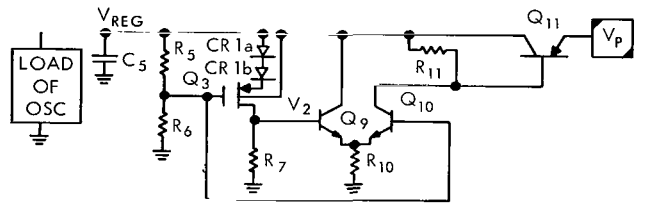
$$V_{MOS}' - V_{MOS} \approx \Delta V_{GS} \left( \frac{\partial V_d}{\partial V_{GS}} \bigg|_{I_d} \right) = \Delta V \left( 1 - \frac{R_6}{R_5 + R_6} \right) (-\mu) = -\mu \left( 1 - \frac{R_6}{R_5 + R_6} \right) \Delta V$$



(a) ORIGINAL REGULATOR CIRCUIT



(b) IMPROVED REGULATOR CIRCUIT



(c) FINAL REGULATOR CIRCUIT

Figure 13—Regulator circuits.

where the amplification factor  $\mu$  is typically about 40; therefore

$$\Delta(V_1 - V_2) \approx + \left(1 - \frac{R_6}{R_5 + R_6}\right) \Delta V + \mu \left(1 - \frac{R_6}{R_5 + R_6}\right) \Delta V = \left(1 - \frac{R_6}{R_5 + R_6}\right) (\mu + 1) (+\Delta V)$$

$$\approx 20 \Delta V \quad \text{if} \quad R_5 \approx R_6 .$$

Thus a small change in the regulator will produce a bigger change in the difference amplifier, resulting in a closed loop gain about 20 times larger in the improved circuit, as shown in Figure 13b. As the oscillator is cooled, the gains of  $Q_9$ ,  $Q_{10}$ , and  $Q_{11}$  decrease, but the gain of  $Q_3$  increases, making the degree of regulation almost constant with temperature. As mentioned previously, the temperature coefficient of the MOSFET is linear and proportional to the current through the device if the gate and drain are connected together. This is still true in this improved circuit, because the difference amplifier holds  $V_1$  very close to  $V_2$ . Figure 14 shows a comparison of the regulator outputs for these two circuits with the output from a zener diode regulator which was directly substituted into the circuit shown in Figure 13a.

The only problem which developed from the "improved" regulator circuit was that  $V_{MOS}$  voltage drop was only about half of the regulated voltage, at the current needed to achieve temperature compensation. Thus when a change in "threshold" voltage occurred in the devices in  $Q_1$ ,  $Q_2$  (Figure 11), and  $Q_3$  (Figure 13), the regulator over-compensated by a factor of 2. This can be shown mathematically as

$$V_1 = V_{REG} \left( \frac{R_6}{R_5 + R_6} \right) = V_{REG} - V_{MOS} = V_2 ,$$

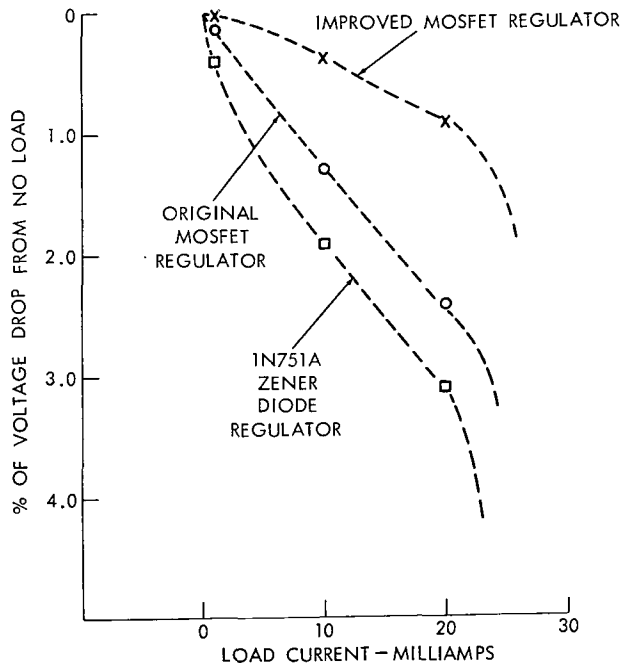


Figure 14—Comparison of regulator performance.

$$V_{REG} = + \frac{V_{MOS}}{\left(1 - \frac{R_6}{R_5 + R_6}\right)} ;$$

where  $R_6 \approx R_5$ ,

$$V_{REG} \approx \frac{V_{MOS}}{1/2} = 2V_{MOS} ,$$

$$V'_{REG} \approx 2V_{MOS} + 2\Delta V_{MOS} .$$

Thus the final regulator circuit, shown in Figure 13c, relieved this problem by three effects, described as follows.

The placement of diodes CR1a and CR1b in the source lead makes the  $Q_3$  MOSFET appear much more like the  $Q_1$  and  $Q_2$  MOSFETs in the oscillator (Figure 11) near the +5 volts input because of the additional source-to-body voltage. These diodes also add a large negative temperature

coefficient to  $V_{REG} - V_2$ , which is canceled out by a large increase in current through  $Q_3$  which also increases the "threshold" voltage of  $Q_3$ , as shown in Figure 15. The result was a voltage reference about 2.5 volts larger for the oscillator, which makes a change in "threshold" voltage of  $Q_1$ ,  $Q_2$ , and  $Q_3$  track more exactly. This is shown mathematically as:

$$V_1'' \approx V_{REG} (.2) \approx V_{REG} - V_{MOS}'' = V_2'',$$

$$V_{REG} \approx \frac{V_{MOS}''}{.8} = \frac{5}{4} V_{MOS}'' ,$$

$$V_{REG}' \approx \frac{5}{4} V_{MOS}'' + \frac{5}{4} \Delta V_{MOS}'' .$$

Thus a change in effective gate-to-source voltage on  $Q_1$  and  $Q_2$  MOSFETs due to temperature, radiation, or charge migration is compensated for by about 5/4 as much change in the regulated supply. The addition of a third diode to increase  $V_{MOS}''$  to almost equal  $V_{REG}$  becomes impractical unless dependence on a negative supply voltage is possible to operate the difference amplifier and supply current to  $Q_3$ . This well-regulated supply is not available in the system herein described; thus we will settle for this approximation.

This final regulator circuit is now able to regulate the voltage supplied to the analog oscillator, and varies almost the same as the MOSFETs within the oscillator as a function of temperature, radiation, and charge migration. The resistors  $R_5$  and  $R_6$  are used to set the +5 volt frequency of the oscillator at room temperature. The temperature coefficient of the oscillator is varied by changing  $R_7$ . The maximum operating current is determined by  $R_{10}$ , while the minimum current and ability to start with power turn-on are determined by  $R_{11}$ . The transients and tendency to oscillate are reduced by  $C_5$ .

The circuit is extremely independent of the supply voltage  $V_P$  because  $Q_{11}$  supplies a current  $\beta$  times that sensed by the difference amplifier  $Q_9$  and  $Q_{10}$ , nearly independent of the collector-to-emitter voltage. The voltage  $V_P$  must be at least 0.7 volt greater than the highest regulated supply requirements.

## THE DIGITAL ACCUMULATOR

The output of the analog oscillator is accumulated in an 8-bit binary accumulator. It is important to understand the sequence of events, so they are illustrated in Figure 16. At  $T_1$ , a new analog commutator input is switched "on," and has until  $T_3$  to complete this switching process. From  $T_3$  to  $T_4$  the analog input drives the voltage-controlled oscillator, and the output frequency is counted. From  $T_4$  to  $T_6$  a new analog commutator input is switched "on." From  $T_4$  to  $T_5$  the

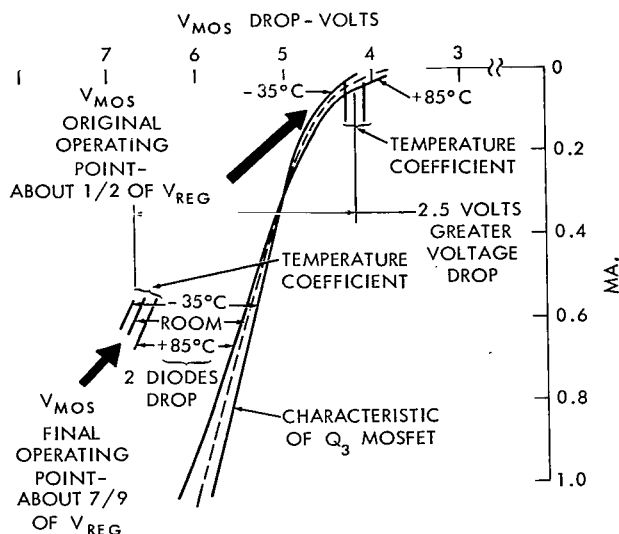


Figure 15—Change in voltage characteristics with additional CR1 diodes.

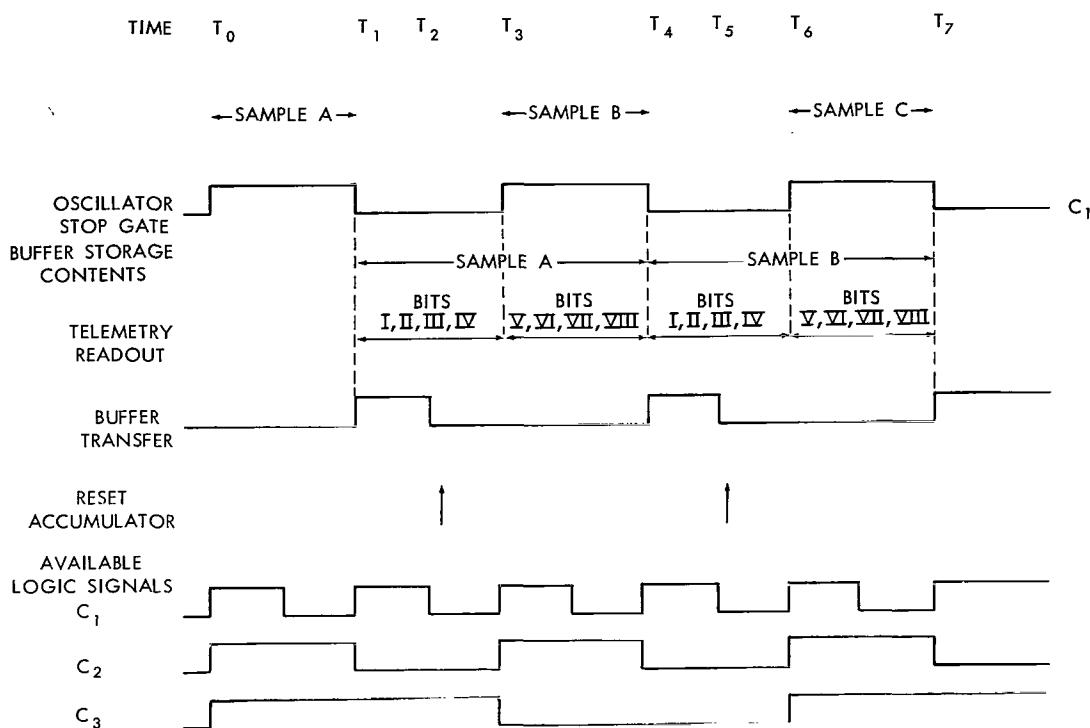


Figure 16—Sequence of events in analog-to-digital conversion.

contents of the accumulator are transferred to a buffer register X. After  $T_5$ , the accumulator is reset and soon is ready to count the next sample.

Meanwhile, the least significant bits of sample B are read out into telemetry. From  $T_6$  to  $T_7$  the most significant bits are read into telemetry while sample C is being digitized.

A block diagram of this circuit is shown in Figure 17. The entire accumulator, buffer storage, and readout logic will be about 350 MOSFETS integrated on one chip within a 22-lead round flat pack.

The system now being fabricated for the IMP I spacecraft uses two analog-to-digital conversion subsystems with a total of 117 analog inputs, and digital storage in 121 separate accumulators. One subsystem makes conversions every 10 milliseconds for 800 milliseconds, and recycles every 2.5 or 10 seconds depending on the telemetry rate. The other subsystem makes a conversion every 40 milliseconds or every 160 milliseconds depending on the telemetry rate.

## THE ANALOG OSCILLATOR SYSTEM PERFORMANCE

The flight-quality analog oscillator performance will be described, along with calibration procedures and testing methods employed. The data shown in Figure 18 demonstrate that the oscillator could be operated and calibrated over a wide range of voltage inputs and frequency outputs. The detailed description of the system performance will be restricted to the circuit to be used in the IMP H, I, and J, but this treatment does not imply that only these voltage and frequency ranges are practical.

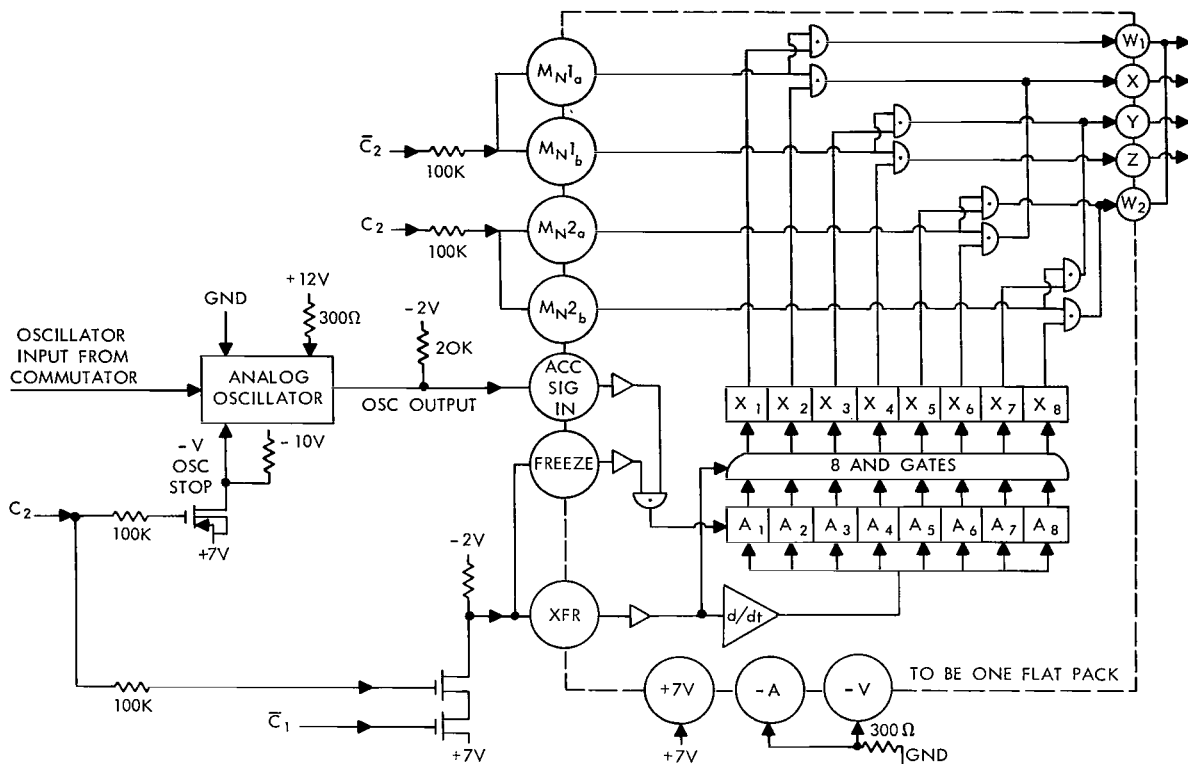


Figure 17—Integrated digital accumulator.

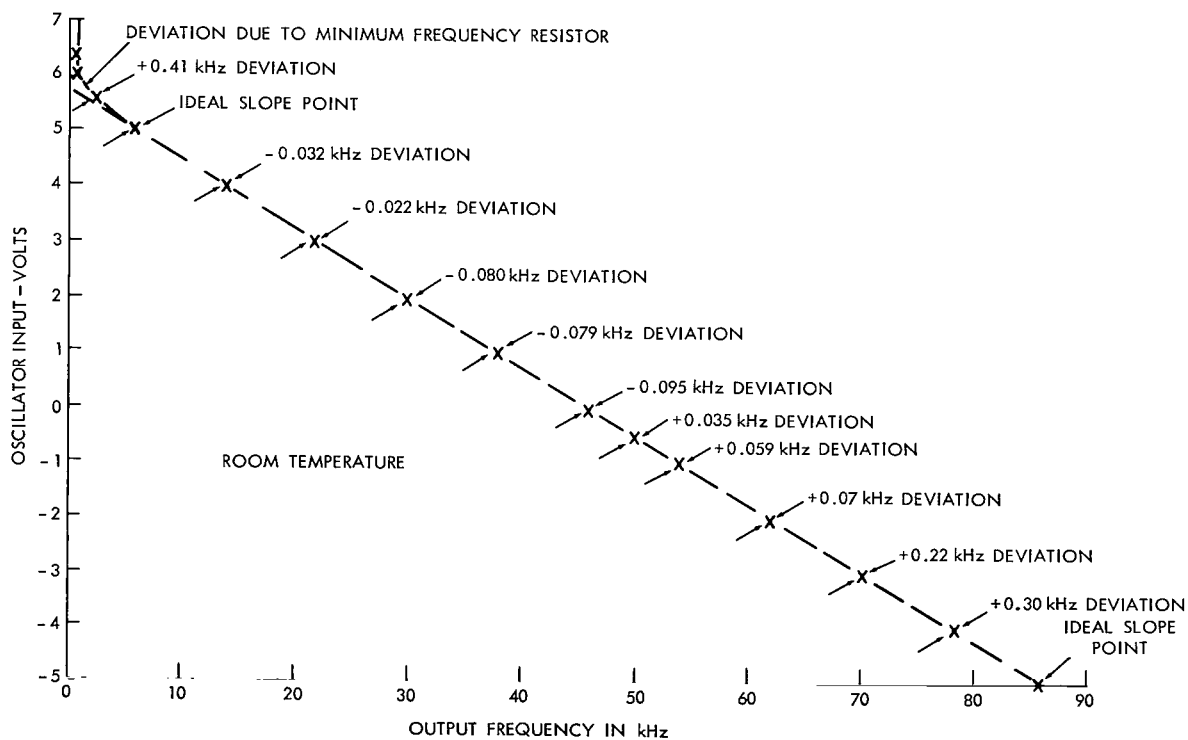


Figure 18—Typical oscillator calibration.

## Test Equipment

The test equipment used to check performance of the analog oscillator is composed of several parts. A set of calibration voltages can be fed into the oscillator manually, or through a flight-quality MOSFET commutator, as previously described. The oscillator can be gated "on" and "off" just as it would be in the flight equipment, with the output counted by a commercial counter and printed at the 3-lines-per-second limit of the printer. The oscillator can be left "on" and the frequency counted to high resolution. Since calibration is needed with greater resolution than flight resolution, the sample time during which the oscillator is gated "on" can be up to 8 times as long as it will be in flight operation. Since the deviation from ideal nominal calibration will be small, only the deviations from nominal will be discussed.

The analog commutator gates into the analog oscillator under test a series of signals, as shown in Figure 19. The first 8 samples are 1-volt increments over the range, with 1/2-volt checks beyond the normal operating range. Samples 9, 11, and 12 are voltages gated into the oscillator in opposite order from Samples 6, 4, and 2 to check for hysteresis effects of the commutator and oscillator. That is, a large input capacitance or local heating of the MOSFETS converting input voltage to current does occur and causes Sample 6 to disagree with Sample 9, Sample 4 to disagree with Sample 11, and Sample 2 to disagree with Sample 12 by about 1/4 of a flight count, or about 0.1 percent.

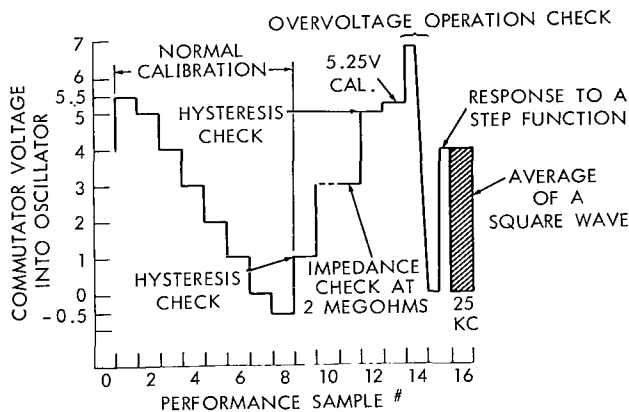


Figure 19—Test equipment commutator output.

The commutator Sample 10 is 3 volts applied through 2 megohms. The slight leakage currents of the MOSFETS, modules, and cables can be detected, and normally produce about a  $\pm 0.05$ -kc deviation from the low-impedance 3-volt frequency. This represents an 8-millivolt change in voltage, or a 2-nanoampere input current to or from the oscillator. This measuring technique is simple but is susceptible to the noise from stray capacitance coupling to the oscillator video output and gating functions.

Sample 14 applies an abnormally high voltage to the oscillator, and then brings the voltage down quickly. Early in the development program, this test demonstrated a problem and, as a result of this test, the minimum frequency resistor was added to the multivibrator to insure operation even with an instantaneous overvoltage. This test continues to check operation during an instantaneous overvoltage condition.

Sample 15 applies a voltage step function to the oscillator input at the same instant the oscillator is gated "on." Any large capacitance effects through the commutator or at the



19



oscillator input would delay the voltage step and produce a much higher oscillator output count if a problem existed. This specifically checks the phased delays of the 100 k ohm input isolation resistors  $R_{2,1}$  and  $R_{2,2}$  which were added to allow operation to continue if either MOSFET (or both) should short because of electrostatic discharge. It is believed this resistance lessens the possibility of such damage and would allow redundant, parallel operation of analog oscillators if desired. The step function is produced by a complementary transistor inverter with a matrix gate signal as its input, and its output shunted to +4 volts.

The last test applies a 25.6-kilocycle phase-coherent square wave to the oscillator input, between nearly ground and nearly +4 volts. If the oscillator performs a true averaging of the sample, then the output should be the same as the 2-volt sample (Sample 5 in Figure 19). The final oscillator circuit square wave average is less than one count (0.5 percent) different from the 2-volt calibration. By contrast, such a response from a high-impedance transistor emitter follower was found impractical because of the unsymmetrical impedance of such circuits. The MOSFET, however, is symmetric with low capacitance and nearly infinite impedance, thus the MOSFET channel current is modulated proportionally to the input voltage, even at a high repetition rate.

## Final Analog Oscillator Circuit Calibration

The final oscillator is shown in Figure 20 with several resistor networks in series and parallel to aid calibration. As shown in this diagram,  $R_7$ , which is trimmed by  $R_{11}$ , allows adjustment of the regulated voltage to make the room temperature frequency at +5 volts nominal. Then  $R_{10}$  is increased to trim  $R_6$  if the 3-volt frequency is higher than nominal, or decreased if lower than nominal. Finally  $R_5$  is trimmed by  $R_9$ , to adjust the oscillator frequency at ground to the desired frequency. These three adjustments interact, but by successive approximation the three points can be adjusted as close to nominal as desired. The oscillator is then temperature-cycled from  $-35^\circ\text{C}$  to  $+85^\circ\text{C}$ . A chart of typical resulting calibration deviation is shown in Figure 21.

If the calibration is generally too high in frequency at high temperature and too low in frequency at low temperature, the voltage regulator temperature coefficient can be decreased by increasing  $R_{12}$  or trimming resistor  $R_{13}$ . A temperature profile of the oscillator is shown in Figure 22 under two different temperature

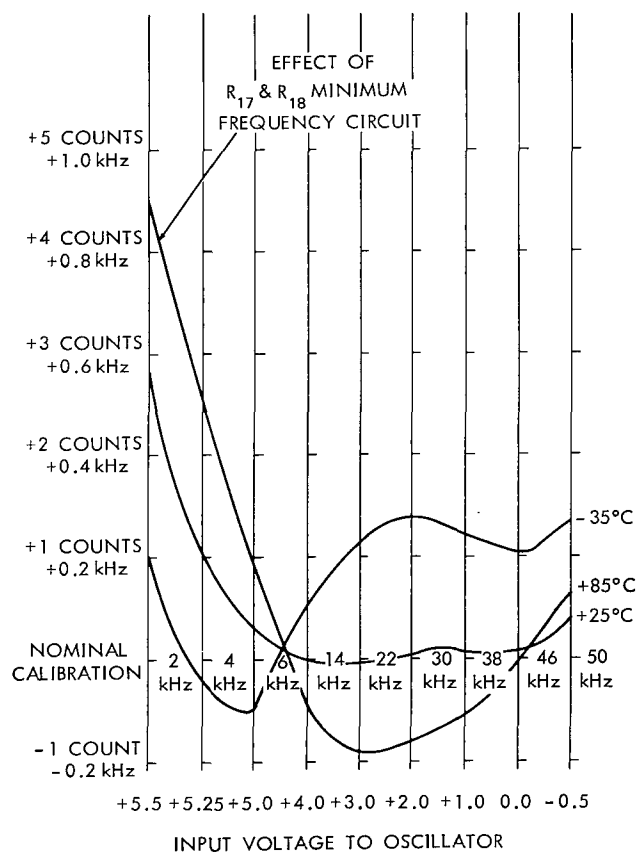


Figure 21—Oscillator calibration deviations from linear ideal.

FIRST CALIBRATION = - - - - - } CHANGED ONLY VOLTAGE REGULATOR  
 SECOND CALIBRATION = - - - - - } TEMPERATURE COEFFICIENT

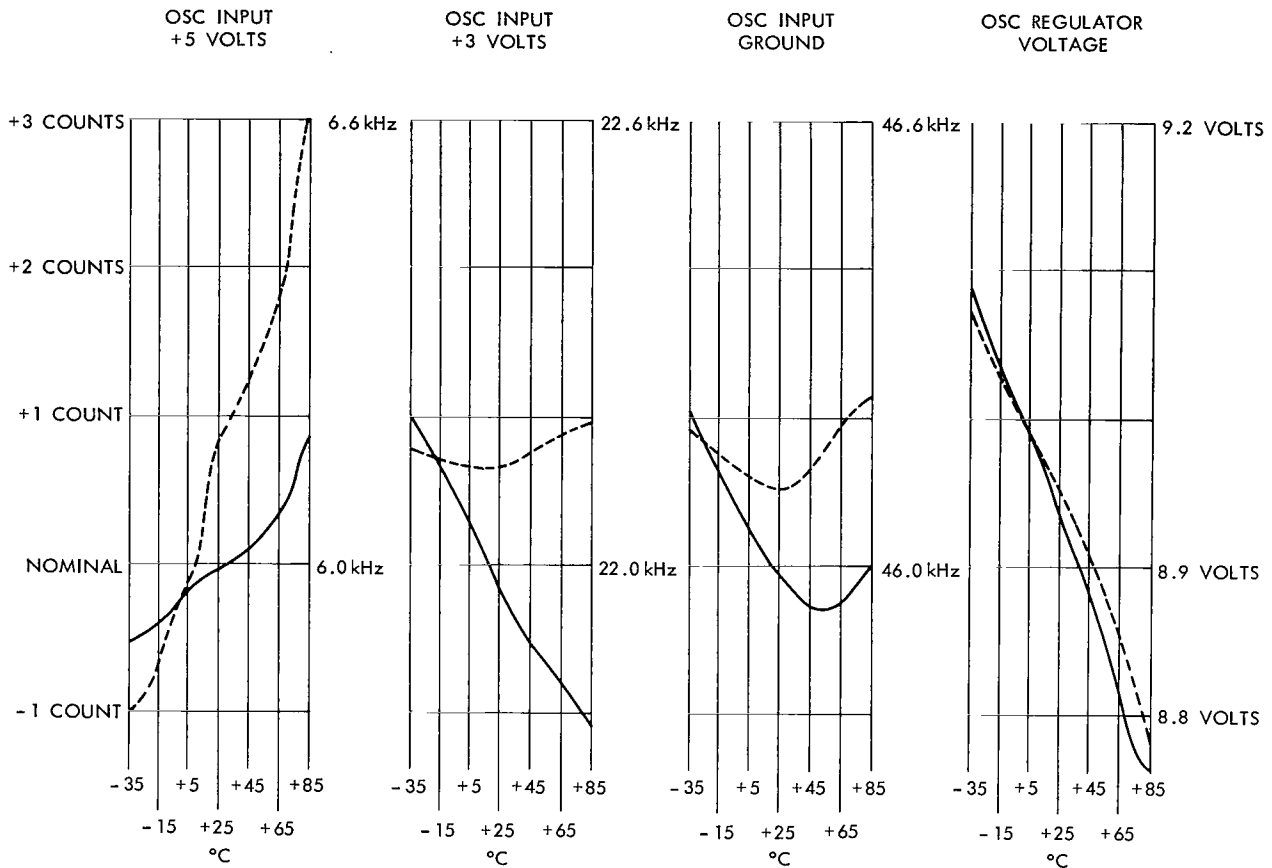


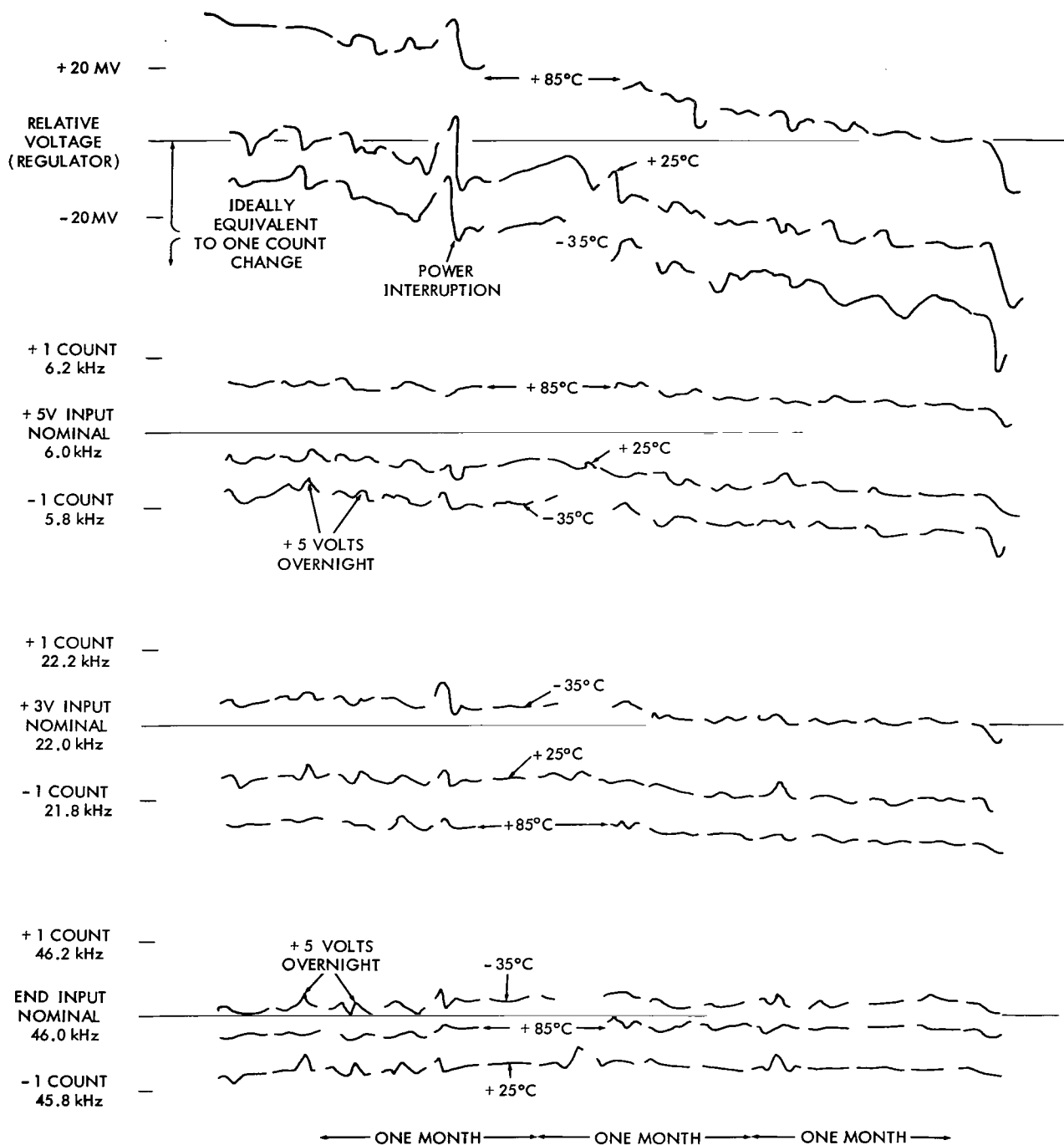
Figure 22—Oscillator temperature profile.

coefficients of the voltage regulator. The positive temperature slope at +5 volts and negative temperature slope at +3 volts are not individually adjustable but are a result of small differences in the temperature coefficient of the MOSFET threshold voltage at different currents. The temperature characteristics with a ground input are also not adjustable, but are affected by the MOSFET in can  $L_1$ . Thus a more favorable temperature coefficient MOSFET can be substituted in the oscillator. These temperature effects are small for our application. Because the characteristics of the MOSFETS, resistors, and capacitors are so similar, consistency between oscillators is quite easy to obtain to within  $\pm 1$  count over the  $-35^\circ$  to  $+85^\circ$  temperature range. The resulting temperature coefficients are within the tolerances needed for this application.

### Long-Term Stability

The long-term stability of the oscillator is verified by the following procedure. Each work day, the room temperature calibration is taken, then the oscillators are cooled for 3 hours

22



covering a period of over three months with about 65 temperature cyclings. The day-to-day repeatability normally appears to be within  $1/5$  of a count, or 0.1 percent, which can be due to a  $\pm 2.5$ -millivolt variation in input voltage. Long-term stability appears to be a function of two effects. One multivibrator was operated with fixed resistors in place of the MOSFETS and experienced a 0.048-kc steady increase in frequency over a three-month period. This represents  $1/4$  count increase after about 65 temperature cycles in three months. The regulated voltage slightly increases in some oscillators, and decreases in others. This probably reflects a change in threshold voltage of the MOSFETS within both the regulator and the oscillator. If the decrease in MOSFET threshold voltage as shown in Figure 23 had not been compensated for by the decrease in regulator voltage, a 2-count increase in the +5-volt calibration should have occurred. But because the regulator slightly overcompensates for the decrease, the frequency drops almost  $1/2$  count at the 5-volt calibration. However, the long-term increase in multivibrator rate causes more of an effect at the ground calibration, increasing the calibration about  $1/5$  count. The MOSFET voltage change affects the ground end less, because such a change is a much smaller percentage of the total frequency-determining current.

The effect of leaving the oscillator at +5 volts instead of the customary ground input overnight affects some oscillators more than that shown in Figure 23. The entire calibration will be as much as 1 count, or 0.2 kHz, higher than normal. The speed with which the calibration returns to normal is temperature-dependent. At  $-35^{\circ}\text{C}$  the calibration is still high by  $1/2$  count after 4 hours. But at  $+85^{\circ}\text{C}$  the calibration returns in a few minutes. This has been found to be due to a change in the MOSFET threshold, and could be some charge migration effect. This effect has been eliminated in flight use by regularly gating a voltage above +6.5 volts into the analog commutator (at a 5-megohm impedance) at a time when no analog-to-digital conversion is required.

Figure 23 shows the long-term stability at the extremes of temperature, with no noticeable variation in temperature coefficient with time. These tests are much more severe both in amount and frequency of temperature cycling than are ever expected in flight. The oscillator appears quite stable even under fast changes in temperature, although this is not important in our application.

The characteristics soon after power turn-on are important. They are tested by automatically commutating the calibration voltages as soon as power is applied, and comparing them with the calibration 5 minutes later. Typically, the frequency drops 4 counts from the previous long-term calibration, and tends to return to this calibration value exponentially; so that it is only about 2 counts lower than normal after 5 minutes, and 0.5 count lower than normal after 24 hours.

These tests indicate that the largest instability will be at the +5-volt calibration, because of the small differences between the regulator and input MOSFET threshold voltages. Since one of the MOSFETS in the voltage-to-current converter affects the +5-volt frequency calibration much more than the other does, the regulator MOSFET is put into the same can with that one, to minimize temperature and radiation gradients.

The effect of changing the input supply voltage to the regulator is less than 0.1 count over a range from +10 to +13 volts.

## Power Dissipation

The power consumed by the various parts of the oscillator is shown in Figure 24. The regulator consumes the major portion of the power because the oscillator load is so small. The total current is monitored during testing to insure proper power dissipation.

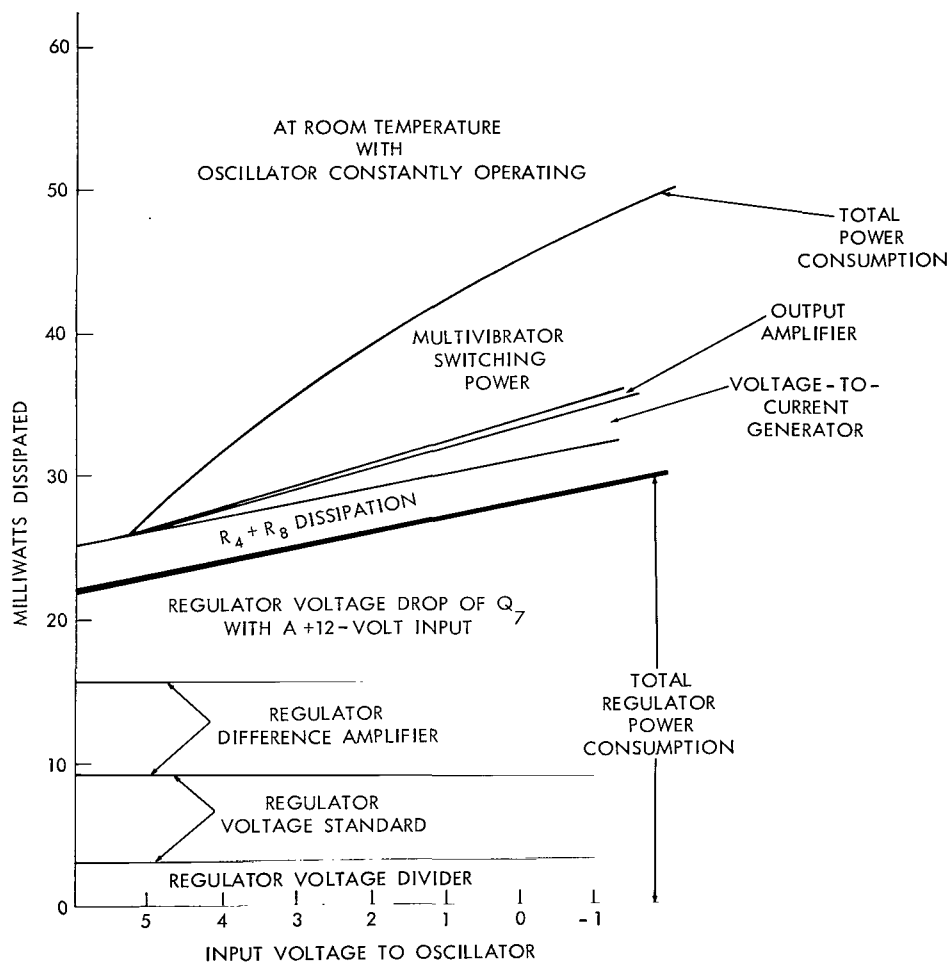


Figure 24—Power consumption breakdown.

## Mechanical Layout

The oscillator is calibrated in breadboard form, then the components are assembled in welded modules which are laid out to minimize capacitive coupling between the high-speed switching circuits and the high-impedance circuits. The change in calibration is about 1 count because of the reduced stray capacitance. The final stage involves potting the module in ECCO FOAM except for

the trimming resistors. After a month's burn-in, final calibration adjustments are made before potting of the trimming resistors.

## SUMMARY OF RESULTS AND CONCLUSIONS

A low-power, low-speed analog voltage commutator has been developed that utilizes MOSFET switches to gate analog inputs sequentially into a voltage-controlled oscillator with no measurable increase in loading during the sample.

The oscillator input employs MOSFET voltage-to-current converters to drive a multivibrator from an internally compensating voltage. A matched MOSFET voltage reference compensates for changes in the MOSFETS due to temperature and radiation. For the full 3-octave frequency range, linearity and stability after turn-on can be held to within  $\pm 1$  percent from  $-35^{\circ}\text{C}$  to  $+85^{\circ}\text{C}$  for inputs from 0 to +5 volts.

A great saving in size and cost of the digital output storage (Figure 25) has been made possible by using a MOSFET integrated circuit with all the storage in one package, thus making multiple sample analog-to-digital conversion data storage practical.

The potential applications for this analog-to-digital conversion system appear quite broad, and may find use where ultra-high input-impedance and microminiaturization are important. The extremely broad range of frequency and voltage over which these concepts could be extended may allow measurements to be made which were impractical previously.

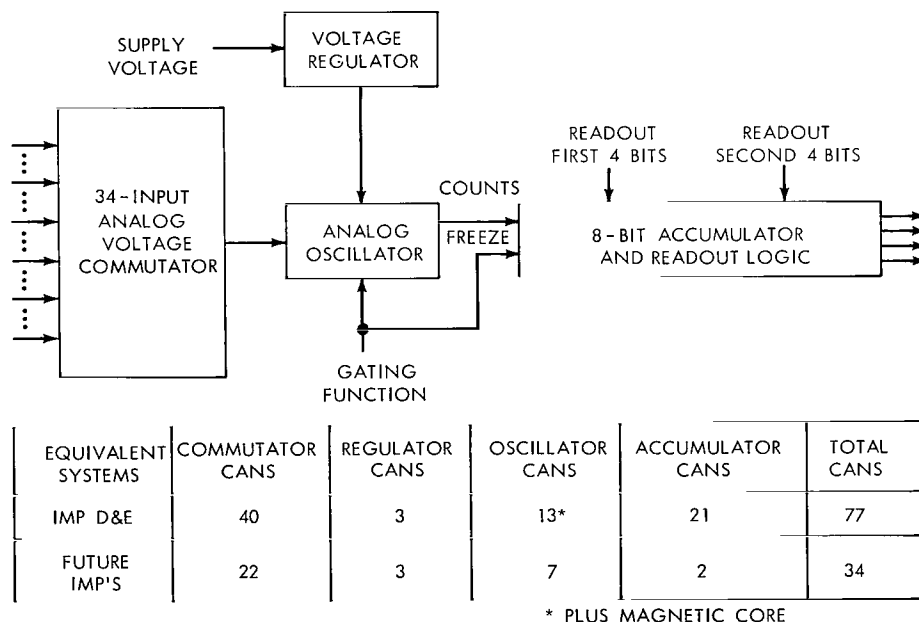


Figure 25—Savings in cans by more highly integrated MOSFET technology.

## ACKNOWLEDGMENTS

The development of this system involved a learning process over a two-year period. I wish to acknowledge the job performed by Kenneth Rehmann in the time-consuming developmental testing phase of this work. The design testing and initial calibration of 20 flight-quality analog oscillators were performed by Carl Dickson. The uniform linearity and temperature characteristics were obtained by Mr. Dickson through the patient selection of the calibration resistors by successive approximations. I wish to thank him for his perseverance and help in gathering the data which made this report possible. Final calibration and testing were done by Howard Holby, confirming the long-term stability of all 20 oscillators.

Goddard Space Flight Center  
National Aeronautics and Space Administration  
Greenbelt, Maryland, October 22, 1968  
861-51-75-01-51

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