MICROMINIATURE ACCELEROMETER/TRANSMITTER

CONTRACT NO. DOT-TSC-1095

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GENERAL

This paper describes a miniaturized nine-channel accelerometer/telemetry system built into an athletic mouthpiece. This system, being developed for the Department of Transportation, telemeters data from the mouthpiece to a receiver/demodulator, using the 90 MHz FM band. A microprocessor, built into the demodulator, computes three angular accelerations, on line, thus providing six principal outputs--three linear and three angular, plus six additional linear channels which may be used for additional, off-line data computation.

Each of the nine accelerometer channels has a range of ±125 g's, and a frequency response of 0.5 to 500 Hz. The angular accelerations have a full scale of radians/sec.^{2}, with alternate scales available.

The design of the system involved five closely interrelated design problems: the accelerometer array, the mouthpiece electronics, the RF receiver, the data demodulator, and the demodulator microprocessor.

ACCELEROMETER

The basic accelerometer array consisted of a central triaxis unit, plus three orthogonally oriented pairs, separated from the origin along each of three orthogonal axes, in the manner specified by King. In addition to the ±125 g range and 500 Hz response required, a 5 mV minimum F.S. output and a 0.56"^{3} size for the nine axis array (NovAcel) was chosen.

The small size of the assemblage and the fact that the individual channel outputs would be used for computation of angular acceleration problem creating restraints. In addition, the low quantity of NovAcels likely to be built, and the probable sporadic requirements for them, imposed assembly design problems.

Considerable analysis of potential error sources was performed, in order to achieve best performance commensurate with small size. For example, because the distance between pairs of accelerometers was small, two errors had to be considered: one, that the great precision was desired in reproducibily establishing interaccelerometer distance, and, two, that errors would be introduced into simplified rotational component calculations if the centers of mass of the three accelerometers at the origin were not coincident. These errors are large if the arms are small, and the errors are not amenable to correction to the limited data processing available to the demodulator.

Because angular acceleration would be computed from linear accelerations, it was important to limit cross axis acceleration response. Two types of cross-axis response, the first, simple bending of the beam in its thick direction, the second, accelerations along the axis of the axis of the beam acting on the mass when it is deflected from rest. The first error is predicted to a first order. The second potential error is a complex function of instantaneous position and acceleration, an effect which is not symmetrical, and which is strongly dependent on the deviation from perfect alignment of the accelerometers. Analysis demonstrated that, to compute rotational acceleration to within a few percent, beams would have to be repetitively aligned to better than 1^{o} of angle, with respect to desired coplanar and orthogonal positions.

A unique design was evolved which provided a common point for the c.g.'s of the three central acceleromters, and precision alignment of all nine accelerometer beams and their masses, without depending on the specific performance of the assembler.

The method chosen was to make each set of three accelerometers from four sheets of precision photo-etched beryllium copper. One sheet forms a spacer, two sheets contain the masses, and one sheet contains the three accelerometer beams. The four sheets are aligned by dowel rods through their alignment holes and cemented together; the seismic masses are then cut loose from their alignment supporting structure. Each of the three accelerometers is perforce coplanar and precisely separated from and angularly aligned with its mates. Three such three-accelerometer sets comprise a nine accelerometer array, with the unique feature that the central three accelerometers of the nine-accelerometer set are of forked construction, and intersect each other in a non-interfering manner so as to have a common c.g.

A block of steel is machined to provide support and alignment for the accelerometer array. The only precision machine work needed is on the block. The three planar arrays are bonded to three front faces of the block, the outer case support is bonded to the three rear faces of the block, a cover is then attached, and the assemblage filled with damping fluid.

The actual length, width, and mass configuration of the beams were arrived at after a computer run which calculated various combinations. The conflicting requirements of high resonant frequency, adequate strain level, small size, clearances between the central intersecting accelerometers, and the constraints of using standard material thicknesses made calculations impractical to perform by hand. The computer was programmed to present several alternates; the configuration chosen was selected on the basis of the best clearances, and hence easiest assembly. At this time, the prototype NovAcel has been completed. A few minor changes in case parts and perhaps an upping of beam resonant frequencies are the only changes anticipated for the final units. A full scale linear shake test to characterize the NovAcel's dynamic behavior is planned for the near future.

MOUTHPIECE ELECTRONICS

Background

The mouthpiece electronics performs three functions: amplification of the nine strain gage signals, encoding them into a single data channel while preserving amplitude and phase data within the bandwidth requirements, and transmission via a radio transmitter from the measurement area to the monitoring point.

The mouthpiece electronics is a sampled data system. In order to transmit nine channels of data over a single radio link, the channels are sequentially sampled on a periodic basis. The minimum requirement of sampling frequency is set by the Nyquist criteria. This criteria is a minimum sampling rate of twice the highest frequency component to be transmitted. Practical limitations of implementing a filter to remove unwanted components from the recovered signal limit the sampling rate to a minimum of 2.5 times the highest frequency component. The relationships of the data, sampled data and filter response are shown in Figure 2.

In order to minimize channel to channel phase shifts near the upper bandwidth limit, all channels are sampled simultaneously, and the signals are stored until sequentially transmitted. This insures there is no sampling related differential phase shift between channels while allowing sequential transmission of the samples.

Implementation

The strain gages are excited by one of the cells of the four cell battery. At full scale of 125 G's, the output of the strain gage is approximately 5 millivolts. Nine capacitatively coupled amplifiers boost this signal up to approximately 0.8 volts. These signals are routed to a nine channel sample-and-hold circuit, which periodically connects the outputs of the amplifiers to the holding capacitors for a short interval. At the end of this interval, each capacitor is holding a charge proportional to the analog level of the channel at the time of the sample. These samples are then sequentially transmitted by an analog multiplexer to the analog-to-pulse-width modulator.

The analog multiplexer actually combines the functions of commutator and sub-commutator. During each cycle, the nine channels are sequentially sampled, then, during the tenth interval, one of the subcom channels representing either ± full scale, ± 1/2 scale, zero or sync is transmitted. Each cycle another of these subcom signals is transmitted in sequence. Although the block diagram shows two separate multiplexers for conceptual purposes, actually a single monolithic multiplexer is used to minimize channel to channel imbalances. The address to the multiplexer is modified by the digital logic to provide the desired sequence.

An amomaly in the operation of the multiplexer chip used was observed during elevated temperature operation of the electronics. During transition from one channel to another they were momentarily shorted together. Because the signal input to the multiplexer are actually just charge levels in capacitors, the momentary connection of sequential channels caused charge to transfer from one to the next. This appeared as a severe inter-channel modulation. Modifying the digital electronics, which produces the addresses for the multiplexer, to switch momentarily to an unused channel (#16) between active channels eliminated this problem. Figure 3 shows the timing relationships. This error has been reduced to less than .5 percent.

The sampled analog signals are then converted to a pulse-width modulated signal. This is accomplished by applying the signal and a linear ramp to a comparator. This is depicted in Figure 4. The resulting pulse train is then used to frequency modulate a low power transmitter operating in the 90 MHz region. Because the modulating signal is either a logical one or zero, the transmitter actually shifts between two frequencies separated by approximately 100 Kc. This frequency-modulated signal is radiated by a small antenna to the monitoring point.

The transmitter has several unique features. Because of the large deviation needed for the bandwidth of information transmitted, and the extremely small power budget, crystal control of the transmitter frequency is not practical. Use of a simple oscillator radiating from the tank circuit as has been used is not satisfactory either, due to severe (greater than 2 MHz) shifts of the operating frequency caused by changes in absorption and dielectric constant of its environment.

A compromise approach was followed, using an LC controlled oscillator operating at one-third of the final frequency, driving a tripler-amplifier which provides isolation. Since the frequency determining elements are physically isolated inside a hermetic container, and electrically isolated by the tripler-amplifer, good stability and large deviations are possible. The improvement in stability over the simple oscillator is approximately two orders of magnitude. (<20 KHz)

Power for the mouthpiece system is provided by four 1.5 volt silver-oxide cells, each approximately 0.4 inches in diameter by 0.15 thick. These cells will operate the system for two hours before degradation of accuracy or transmitter output occurs.

RECEIVER-DEMODULATOR

The receiver used is a modified commercial FM tuner. Tests with an expensive, high quality telemetry surveillance receiver indicated that its features of variable bandwidth, etc. did not provide any better performance than the commercial unit. The 220 KHz bandwidth of the commercial unit matches the nominal 100 KHz deviation of the transmitter well, and demodulates the signal with minimum distortion and ringing. A relatively simple conditioning circuit restores the zero crossing points and provides a clean signal to drive the demodulator.

The demodulator is digitally based, using a reference oscillator and counting its output for varying intervals corresponding to the pulse width of the waveform representing each of the channels. This is shown in Figure 5. The reference oscillator is part of a phase-locked-loop which maintains its frequency at precisely 1024 times the PWM rate. This phase lock loop compensates for possible drifting of the mouthpiece clock oscillator, which would otherwise appear as an offset in all channels. As shown, a single counter of 10 bits (corresponding to 1024) counts the reference oscillator during the time defined by a logical one for each channel. At the time the PWM signal goes low, the count in the counter represents the analog signal level transmitted. This number is entered into the microprocessor by the "data ready" strobe for storage and further processing. The counter is then cleared (set to Ø) and waits for the next channel sample to begin. A "channel 1" signal, occurring simultaneously with the "data ready" signal for channel 1, identifies the data channels to the microprocessor since their order is fixed. This format of ten leads carrying the data and a "data ready" signal is also used at the output of the microprocessor.

The outputs of the demodulator are routed to 9 ten-bit latches. They in turn feed D to A converters which generate an analog output voltage. These voltages are filtered by 9-pole Butterworth active filters. These filters have a -3dB roll-off point of 500 Hz, and a -54dB/octave roll-off rate. This results in an 85 dB attenuation of the unwanted frequency components.

Scaling of the output level is accomplished by a combination of varying the reference voltage to the D to A converters and shifting the ten data lines to obtain scaling by factors of two.

Subcom data is recovered in the same manner as signal data, except the counter is enabled once for every six cycles through the channel sequence. The position relative to the sync signal (which is a missing pulse) defines which of the five subcom signals is being decoded. Since this data is only a slowly changing DC calibration level (mV/hr) a simple filter with an upper cutoff of 10 Hz is sufficient to remove the unwanted noise. These calibration signals are also used by the microprocessor (described below) and are presented in the same format as the data channels.

MICROPROCESSOR

The function of the microprocessor is to compute rotational acceleration from the nine linear accelerations and the lengths of the moment arms, provide automatic zero-offset correction, and provide programmable sensitivity scaling (calibration) of the channels.

The processor is a 16-bit bipolar unit, having an execution time of 100 nsec. It is comprised of eight 2-bit slices, with the appropriate support chips. It can perform a 16-bit by 16-bit multiply in approximately 15 μsec., and is micro-programmed for maximum speed.

The algorithm which is performed to extract the rotational acceleration is of the form:

$$\dot{\omega}_X = \frac{(A_{Z_1} - A_{Z_0})}{2\,\rho_{y_1}} - \frac{(A_{y_3} - A_{y_0})}{2\,\rho_{y_3}} \qquad \text{(from King, et. al.)}$$

Rearranging to a more convenient form for the microprocessor and allowing for different values of ρ along each axis:

$$\dot{\omega}_X = \frac{A_{Z_1}}{2_{\rho_{y_1}}} - \frac{A_{Z_0}}{2_{\rho_{y_1'}}} - \frac{A_{y_3}}{2_{\rho_{x_3}}} + \frac{A_{y_0}}{2_{\rho_{x_3'}}}$$

Adding a scaling constant (value of G) to result in proper units, and consolidating constants:

$$\dot{\omega}_X = A_{Z_1} K_1 - A_{Z_0} K_2 - A_{y_3} K_3 + A_{y_0} K_3$$

These constants are computed from the measured separation of various accelerometer pairs, the value of G, and the factor of 2 from the original equation.

This algorithm must be performed three times, one for each rotational acceleration. Each calculation involves four multiplications and four additions. This takes approximately 65 μsec for each calculation, or roughly 200 μsec out of the 800 μsec available for the calculations.

In addition to the rotational extraction, scaling of the sensitivity of each of the nine accelerometers and correction for the zero offset of the input electronics and multiplexing gates for each of the nine input channels is provided. These represent an addition and a multiplication each, or about 144 μsec total. This leaves approximately 50% of the available calculation time available for future expansion or more complex data manipulation.

The calculated constants used in the rotational acceleration calculation, and the nine measured constants for sensitivity scaling are stored in a programmable read-only-memory. This particular memory is paired with the mouthpiece and is its calibration. The zero offset correction is accomplished by a special routine in the microprocessor. After placing the mouthpiece at rest for a long time relative to the low frequency cutoff of the system (0.5 Hz), each of the signals from the nine accelerometers will drop to zero since they are each capacitatively coupled. By the operator's manually closing a switch, the microprocessor is caused to go to a special routine which measures the residual output of each channel. These measurements are stored in the microprocessor's memory, and during normal operation are subtracted from the data to correct for the zero offset.

In order to maintain the small phase errors accomplished by the sample and hold circuitry in the mouthpiece, it is also necessary to output the digital data to the filters simultaneously. The microprocessor outputs all twelve channels (nine linear and three rotational) in less than 6 μsec, thus insuring that the final output signals retain their correct phase relationships.

PERFORMANCE

At this time the entire system has been operated, less only the microprocessor which is currently being mated to the demodulator. Performance has been extremely good.

The total residual noise from all sources is approximately 15 millivolts peak-to-peak, with the channel set up for 3 volts peak-to-peak output full scale. This corresponds to a noise level of 0.5%. The errors are contained within this band, since they correspond to roughly ±1 bit or .1%. Interchannel cross modulation, (except for the momentary simultaneous open gate condition previously discussed) is unmeasurable. Stability versus temperature is excellent:, worst measured zero offset of the inputs electronics over 0^{o} to 70^{o}C was 11%, well within the 25% correction range of the calibration system. Gain stability was better than 1%. The drifting clock rate was less than 1%, also easily accommodated by the ±20% range of the phase locked loop.

CONCLUSION

The system described is a practical and accurate means for remotely measuring linear and rotational accelerations. Techniques developed for this program of electrical design, mechanical design and packaging make it state-of-the-art technology in terms of its small size and high accuracy.

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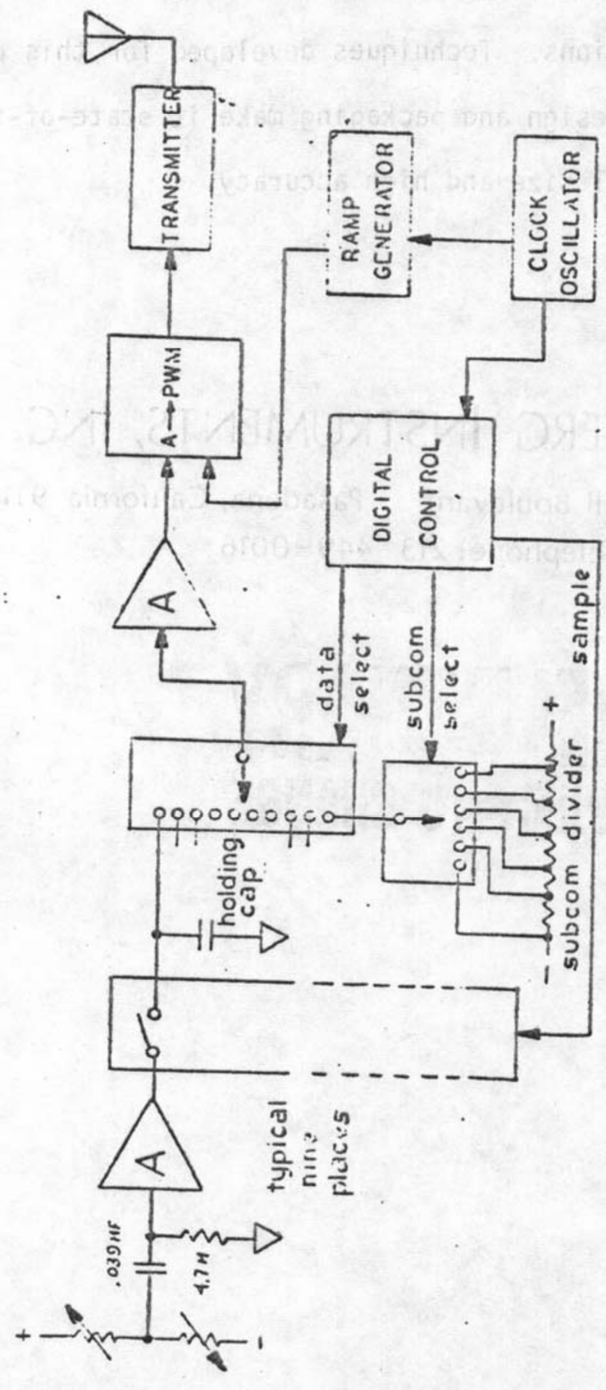
PRELIMINARY
SUBJECT TO CHANGE



MOUTHPIECE ELECTRONICS - BLOCK DIAGRAM

FIGURE A

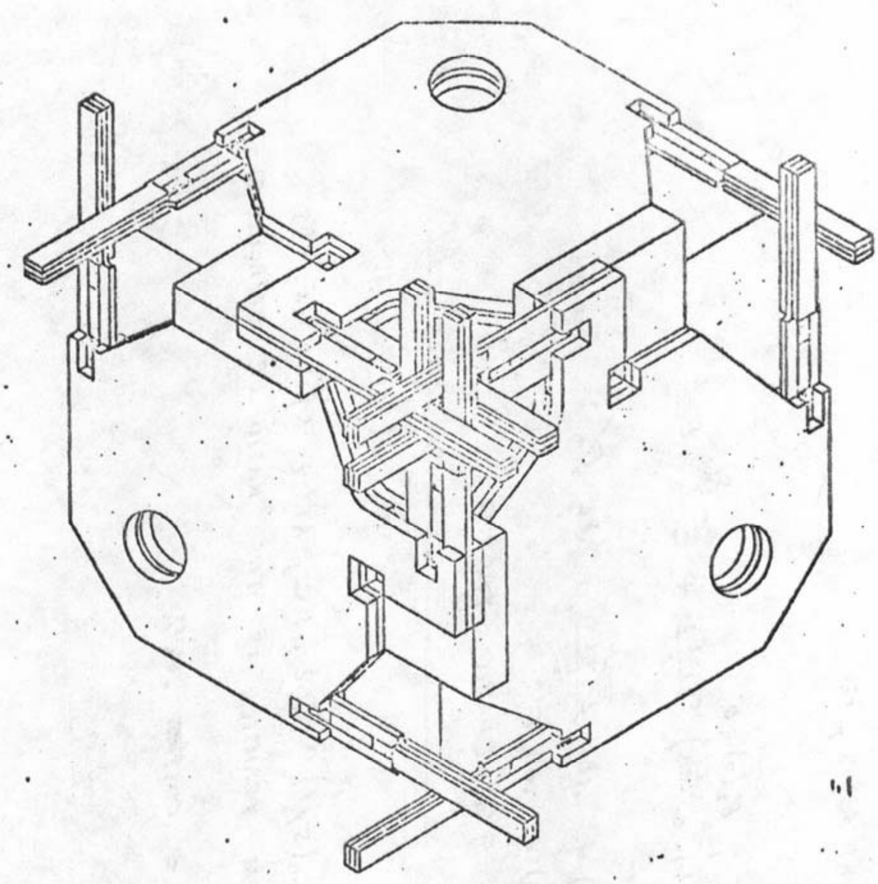
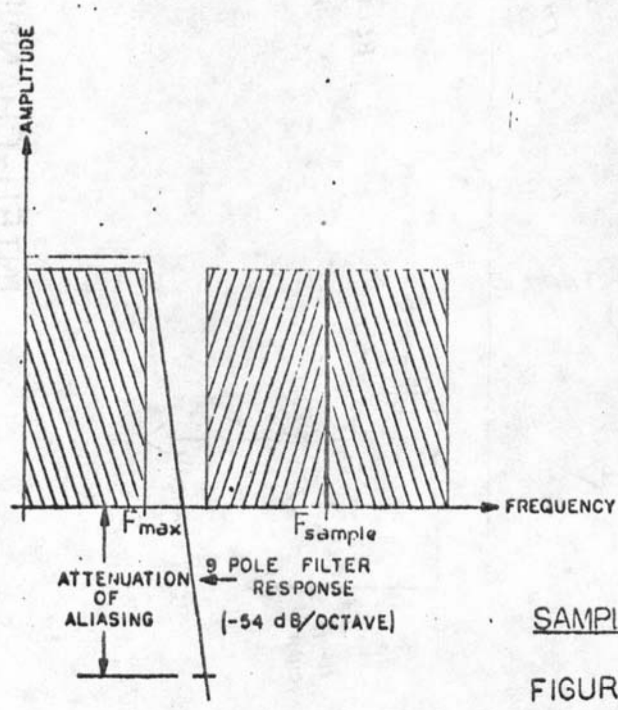


FIGURE 1



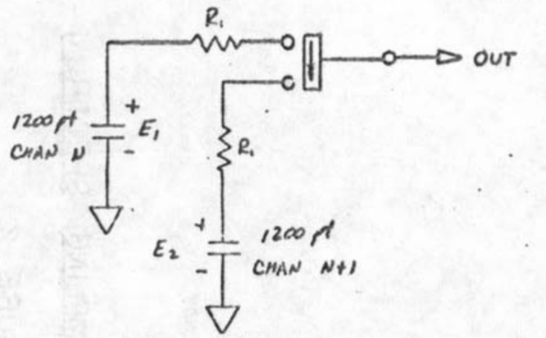
SAMPLING SPECTRUM

FIGURE 2

R_1 - INTERNAL IMPEDANCE OF MULTIPLEXER ($\cong$ 4000 Ω)

R_1C = 10 μ SEC

TIME OF OVERLAP $\approx$ 2 μ SEC



LET $E_1(0) = E$; $E_2(0) = 0$

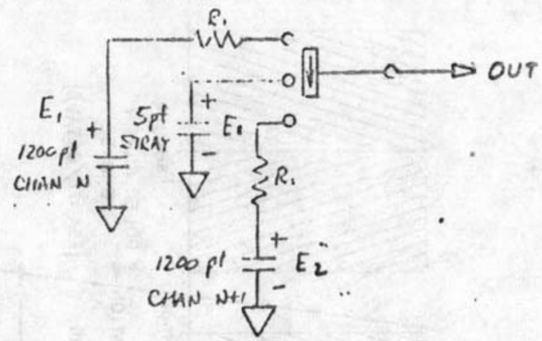
THEN $E_2(t) = \left(1 - e^{-\frac{t}{2R_1C}}\right) \frac{E_1 + E_2}{2} + e^{-\frac{t}{2R_1C}} E_1$

FOR $t = 2\mu$ SEC

$E_2(2\mu\,\text{SEC}) \doteq .18\ \frac{\cancel{E_1} + E_2}{2} + .82\ \cancel{E_1}$

$E_2(2\mu\,\text{SEC}) = .09\ E_1$

$\therefore$ THERE IS 9% CHARGE TRANSFER



BECAUSE $R_1(5pf) \approx .02\ \mu$SEC, AT $t = 2\mu$ SEC $E_S = E_1$

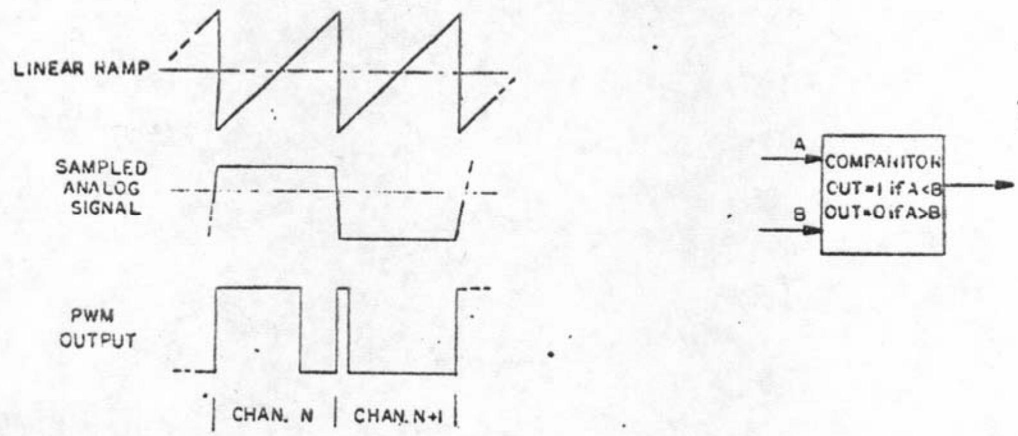
$\therefore$ NET CHARGE TRANSFER IS THE RATIO OF CAPACITIES

$$\frac{5pf}{1200\ pf} = .0042 \equiv .42\%$$

INTER-CHANNEL MODULATION FIX

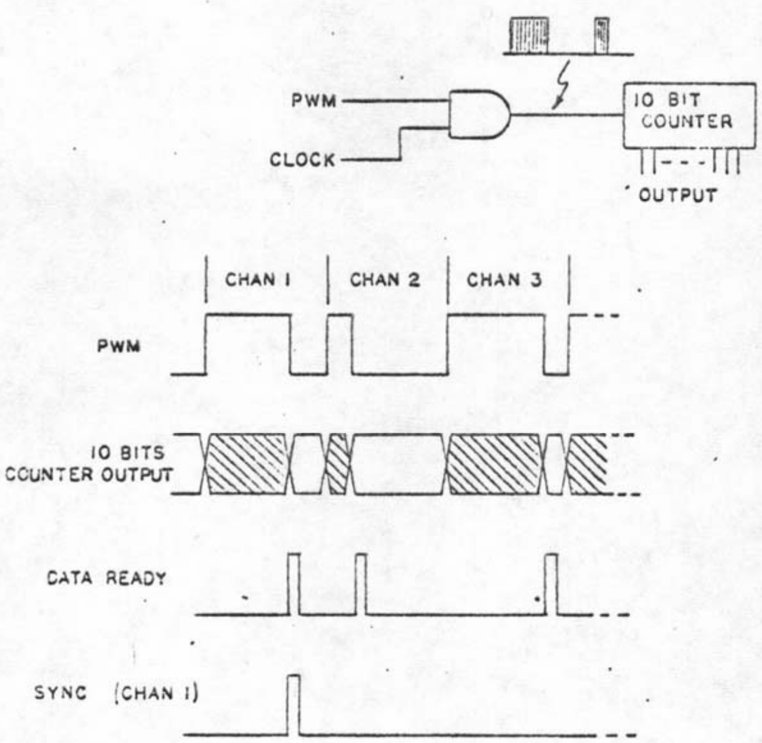
FIGURE 3

135
136?



ANALOG TO PULSE WIDTH CONVERSION

FIGURE 4



DEMODULATOR

TIMING

FIGURE 5