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AN EXPERIMENTAL DELTA-MODULATION SYSTEM

W. Ghurahoo, K. Saunders* and J. N. Tripathi*,*

ABSTRACT

Signal approximation, quantization noise, slope overload noise, and signal reconstruction in delta-modulation are explained. A linear delta-modulation system for speech signals is designed constructed and tested. Quantization noise is experimentally measured and plotted as a function of sampling frequency. Speech signals were clearly reproduced. Experimentally measured signal-to-noise ratio performance is close to the theoretical signal-to-noise ratio expression of delta-modulation.

1.0 INTRODUCTION

Digital transmission of electrical signals has many advantages over conventional analog transmission. Digital signals can be transmitted over longer distances with very little loss of quality and with greater probability of correct reception. Digital signals can be time-division multiplexed unlike the analog signals which are frequency-division multiplexed. Therefore, complex multiplex filtering associated with frequency-division multiplexing can be avoided since time-division multiplexing can be accomplished by digital circuits.

Differential pulse-code modulation (DPCM) reported in 1950 combines pulse-code modulation and delta-modulation(DM). The PCM system based on 1947 technology was large, rather unreliable and tended to get very hot. In fact the key invention of the transistor in Bell Labs altered this situation. By 1957 the transistor appeared as near ideal switching device, small, very fast, reliable and with very low power consumption. By this time need was felt for a digital transmission system. The Bell T1 system proved to be highly successful and is now in widespread use throughout the world. This, in fact, represents the starting point for modern digital communications.

It is difficult to make a definitive comparison between the performance of PCM and DM because DM technology is continuously improving. If the bit rate can be made arbitrarily large, then PCM will outperform DM. But if the systems are compared for successively lower bit rates than a crossover will be reached below which DM gives better performance than PCM [2].

Most commercial satellite links to date use only PCM for digital transmission while DM has been used for military applications or specialized civil systems like the space shuttle in which the simpler equipment of delta-modulation and its improved performance advantage at low bit rate (i.e., narrow bandwidths) over PCM is significant [2].

This paper presents design and construction of a delta-modulation system using discrete components and discusses its performance for band-limited speech signals. Delta-modulation is particularly suitable for telephone signals [3].

2.0 DELTA-MODULATION SYSTEM

Delta-modulation is the simplest analog-to-digital (A/D) converter and produces two level voltages at its output. The principle of delta-modulation is illustrated in Figure 1. The input message signal is bandlimited by passing through a band-pass filter to obtain telephone quality speech signals. The message signal X(t) is compared with a stepwise approximation $\tilde{X}(t)$ and the difference signal $Y(t) = X(t) - \tilde{X}(t)$ is quantized into two levels $\pm\Delta$ depending on the sign of the difference [4]. The output of the quantizer is sampled to produce

$$Y_{sq}(t) = \sum_{k=-\infty}^{\infty} \Delta sgn[X(kT_s - \hat{X}(kT_s)]\delta(t - kT_s) \tag{1}$$

where T_s is the sampling period.

$$sgn(t) = \begin{cases} 1, & t > 0 \\ -1, & t < 0 \end{cases}$$

The stepwise approximation X(t) is generated by passing the above waveform given in equation [1] through an integrator and a low-pass filter. The integrator responds to an impulse with a stepwise [4]. Since there are only two possible impulse weights in $Y_{sq}(t)$ this signal can be transmitted using a binary waveform. The integrator can be a single integrator, double integrator or any other signal processing network [5]. The demodulator consists of an integrator and a low-pass filter (Fig. 1).

Department of Electrical and Computer Engineering, The University of the West Indies

When X(t) is changing, $\tilde{X}$(t) and $\hat{X}$(t) follow X(t) in a stepwise manner as long as successive samples of X(t) do not differ by an amount greater than the step size Δ. When the difference is greater than Δ, $\hat{X}$(t) and $\tilde{X}$(t) can no longer follow X(t) and a condition known as slope overload occurs (Figure 1).

Assume $X(t) = A\cos 2\pi f_x t$. Then the maximum signal slope is

$$\left[\frac{dX(t)}{dt}\right]_{max} = A2\pi f_x \qquad (2)$$

The maximum sample to sample change in the value of X(t) is $A\,2\pi f_x T_s$. To avoid slope overload this change has to be less than Δ,

$$2\pi f_x A\ T_s < \Delta \qquad (3)$$

or peak signal amplitude at which slope overload occurs, is given by

$$A = \frac{\Delta}{2\pi}\frac{f_s}{f_x} \qquad (4)$$

where $f_s = 1/T_s$ is the sampling rate of the delta modulator.

Slope overloading can be alleviated by filtering the input signal to limit the maximum rate of change by increasing the step size and/or the sampling rate [4].

The quantization noise power falling in message bandwidth f_x is given by [4].

$$N_q = \left(\frac{\Delta^2}{3}\right)\left(\frac{f_x}{f_s}\right) \qquad (5)$$

The signal power S is $A^2/2$. To avoid slope overload A is equal to $\Delta f_s/2\pi f_x$. Therefore, signal-to-quantization noise ratio in delta-modulation under ideal conditions is given by

$$\frac{S}{N_q} = \frac{3}{8\pi^2}\left(\frac{f_s}{f_x}\right)^3 \qquad (6)$$

Apart from quantization noise, in an actual delta-modulation system, there would be other types of noise such as channel noise, and some slope overload noise; therefore the performance degrades and the signal-to-noise ratio (SNR) performance can be represented by

$$\frac{S}{N} = \frac{3}{8\pi^2 D}\left(\frac{f_s}{f_x}\right)^3 \qquad (7)$$

where D is the degradation factor by which the actual delta-modulation system performs below an ideal delta-modulation system.

The quantization noise decreases inversely with the sampling frequency and doubling the sampling rate reduces the quantization noise power by 3 decibels. The signal-to-noise ratio performance of delta-modulation varies with the third power ratio of sampling to message frequencies (f_x/f_s), and doubling this parameter (f_x/f_s) improves the performance by 9 dB.

Reducing the step size Δ reduces the quantization noise but a small value of Δ results in slope overload noise. An optimum size Δ must exist and that size depends upon the characteristics of input signal X(t), and the sampling frequency. Linear delta-modulation (LDM or simply DM) uses a fixed step size and the schemes where the step size and/or the sampling frequency are varied, are known as adaptive delta-modulation (ADM) systems.

3.0 SYSTEM DESIGN AND CONSTRUCTION

Based on the block diagram of Figure 1, the complete circuit diagram of the delta modulator and the demodulator is shown in Figure 2. The various building blocks are described in the following sections.

3.1 Input Bandpass Filter

The input bandpass filter is used for filtering the message signal to limit its frequency components in a range of 300 Hz to 3400 Hz. Delta-modulation is incapable of transmitting dc components, its dynamic range (amplitude A) and SNR are inversely proportional to the message signal frequency f_x (equation 4) and the integrator at the receiver causes accumulative errors [5]. Delta-modulation is suitable for speech signals which do not have dc component and have less energy in higher frequencies [5].

The input bandpass filter is realized by cascading two second order low-pass sections and two high-pass sections. It consists of fourth-order Butterworth bandpass filter with cutoff frequencies of 300 Hz and 3400 Hz. It is shown in Figure 2. The associated Butterworth polynominal [6] is

$$(s^2 + 0.765s + 1)\,(s^2 + 1.848s + 1) \qquad (8)$$

Transfer function [6] of a section shown in Figure 3 is given by

$$\frac{V_o}{V_i}=\frac{Z_3Z_4A_v}{Z_3(Z_1+Z_2+Z_4)+Z_1Z_2+Z_1Z_4(1-A_v)} \quad (9)$$

where the gain $A_V=\dfrac{R_1+R'_1}{R_1}$

For a low-pass section

$Z_1 = Z_2 = R_L$

$Z_3 = Z_4 = 1/C_Ls$

For a high-pass section

$Z_1 = Z_2 = 1/C_Hs$

$Z_3 = Z_4 = R_H$

The transfer function for a Butterworth second-order low-pass filter [6,7] is given by

$$A_{vs}=\frac{A_v}{\left(\dfrac{s}{w_o}\right)^2+2k\left(\dfrac{s}{w_o}\right)+1} \quad (10)$$

where $f_o = w_o/2\pi$ is the 3 dB cutoff frequency and 2k is the coefficient of the Butterworth second order transfer function.

Equation [10] can be reduced to the form

$$\frac{V_o}{V_i}=\frac{A_v}{(R_LC_L)^2s^2+(3-A_v)R_LC_Ls+1} \quad (11)$$

Thus equating the coefficients

$w_o = 1/R_LC_L, \quad 2k = 3 - A_V$

Using standard values of components calculated values are shown in Figure 2.

For the high-pass section, the transfer function [6,7] becomes

$$\frac{V_o}{V_i}=\frac{s^2R_H{}^2C_H{}^2A_v}{s^2R_H{}^2C_H{}^2+sR_HC_H(3-A_v)+1} \quad (12)$$

The Butterworth high-pass second order transfer function [6,7] is

$$A_{vs}=\frac{(s/w_o)^2A_v}{(s/w_o)^2+2k(s/w_o)+1} \quad (13)$$

where $f_o = w_o/2\pi$ is the 3 dB cutoff frequency and 2k the coefficient of second order Butterworth transfer function. Thus

$$w_o=\frac{1}{R_HC_H}, \quad 2k = 3 - A_V$$

Using standard component values, the calculated values are shown in Figure 2.

3.2 Difference Amplifier/Summing Amplifier/ Adder

The difference amplifier provides an error voltage which is the difference between the input signal and the constructed (estimated) signal. The operational amplifier with positive and negative input voltages can provide the amplified difference signal. The operational amplifier (Op Amp), thus acts as a summing amplifier or an adder with proper polarity since subtraction is the process of adding two opposite levels/voltages/ numbers. The slew rate of 741 Op Amp was slow and sampling took place on rising edge of the pulse waveform, hence pulses of widely varying amplitudes were produced. Instead NF 5534 Op Amp was used only in adder circuit due to its high cost and 741 Op Amps were used in all other applications requiring operational amplifiers.

3.3 Quantizer and Sampler

The J-K flip-flop 7473 and a sampling I.C. unit LF 398 perform the quantization and the sampling operation (Figure 2). The 'can' type I.C. LF 398 acts as a two-channel switch. When the input to terminal 8 is high, input at terminal 3 is connected to the output and when this input is low, the input at terminal 8 is connected to the output. The output of the adder is $\pm$ 5 volts, since the adder has a high gain of 100. This was converted to $\pm$ 5 volts or zero voltage levels by using a 1-k potentiometer. To generate pulses with a fast rise time and to provide interfacing, the potentiometer output was passed through two NOT gates (7404) and then connected to the J and the K inputs. Due to the configuration used the J-K flip-flop output was allowed to change only during the positive rise of the clocking pulse generated by the voltage controlled oscillator (VSO) and this is achieved by using two LF 398 units and NOT gates. The additional two NOT gates used before the triggering terminal 8 on second sampler are damping the oscillations of VCO output pulses.

3.4 Voltage-Controlled Oscillator (VCO)

The VCO implementation utilized a 74123 Dual Retriggerable Monstable Multivibrator [8]. The 74123 chip is connected as a square-wave generator, the external resistor pins 15 and 7 are connected to a voltage controlled current source. This current source is essentially a pnp transistor so biased that by varying the base voltage the collector current can be varied. In Figure 2, the variable transistor current source is replaced by a variable 10k resistor tied to the supply voltage of +5V. By varying the potentiometers, the frequency and duty cycle of the VCO can be changed from 10 kHz to 200 kHz.

3.5 Integrator And Low-Pass Filter

The output of the quantizer and sampler is passed to the integrator and low-pass filter (Figure 2). The integrator and the low-pass filter reconstruct the message signal from the delta modulator output pulses. The resistor in parallel with the capacitance in the feedback path prevents the integrator from going in saturation from small input d.c. voltages [7]. Using the design, the error is reduced to less than one percent in comparison to normal integrators. Component values are shown in Figure 2.

The 3 dB point of the low-pass filter occurs where the voltage gain falls by half of the DC gain. The low-pass filter is designed for cutoff frequency of 3700 Hz. The component values are shown in Figure 2.

3.6 Demodulator

The demodulator consists of an integrator followed by a high-pass filter and a fourth-order Butterworth low-pass filter. The demodulator performs the inverse operation of the modulator. The integrator in the demodulator is exactly the same as used in the feedback path of the modulator. Integrator with Op Amp provides voltage gain also and the operation is linear. The high-pass filter following the integrator removes the DC gain on the output of the integrator due to oversampling. This DC bias appeared on the output pulses of the modulator when the sampling clock frequency of the VCO was too high. This means that at such clocking rates, the falloff time of the sampling unit was not small enough to discharge to zero. Thus the output of the quantizer between the pulses is a fraction of a $\pm$ 5V rather than being 0V.

The high-pass filter has a variable gain and it is possible to match the input signal amplitudes. Another potentiometer was included between the input bandpass filter of the modulator and the quantizer unit. This helped to match the signal amplitudes. The highpass filter has a 3 dB cutoff frequency of 159 kHz. The calculated values of the components are shown in Figure 2. The fourth-order Butterworth low-pass is the same as used in the delta modulator.

4.0 SYSTEM PERFORMANCE AND DISCUSSION

Based on the circuit diagram of Figure 2 the complete delta modulator and demodulator was constructed. The modulator and the demodulator worked successfully. The waveforms obtained on the various points are shown in Figure 4. The voltage-controlled oscillator provides a variable frequency output used to sample the quantized difference signal. The VCO output waveform is shown in Fig 4 A,B. At higher sampling frequencies the square-wave waveform has some spikes (ringing oscillations). At lower sampling frequencies the spikes become negligible (Figure 4B). The spikes on the waveform do no in any way affect the circuit operation. (Figure 4, F, G, H).

Idle channel noise is found by grounding the input terminal of the modulator and observing the demodulator output. Idle channel noise waveforms at various frequencies are shown in Figure 4. Figure 4C shows idle channel noise at very high sampling frequency (238 kHz). The upper waveform shows the demodulator output whereas the lower waveform shows the modulator filter output. Figure 4D shows the demodulator (upper waveform) and the modulator (lower waveform) filter outputs at 100 kHz sampling rate. Note that the spikes have disappeared because the VCO output at this frequency does not have any spikes.

As seen from Figure 4 idle noise consists of a series of positive and negative pulses. Because of errors, two pulses of same polarity are often received. Since one step size is greater than another, then for a succession of a positive and a negative pulse, the output is not able to rise (or fall) as much as it should be able to fall (or rise). Hence the output drifts away from ground or 0 V value. Therefore, after a number of alternating pulses the rise (fall) is not able to meet the 0 V input and so another pulse of same sign is required (Figure 4E)

The overall performance of the delta modulator and the demodulator was judged by using audio signals. The audio signals were broadcast signals obtained from an FM receiver. The actual waveforms are shown in Figure 4F. The upper waveform shows the input to the modulator and the lower waveform shows the demodulator audio output. At low sampling frequencies (below 50 kHz) the audio signals were not audible. As the sampling frequency was increased, the audio signals began to be more audible. At 100 kHz the audio signals were clearly heard. At very high sampling frequencies (200 to 240 kHz) the audio output was very clear. The performance of the input bandpass filter and the output low pass filter is shown in Figure 5. The output filter has a sharper cutoff characteristic at higher frequencies. The input bandpass filter has much more symmetrical response both at higher and lower frequencies. The slope of the filter is approximately 12 dB per octave.

Idle channel noise is found by grounding the input terminal of the modulator and by observing the output of the demodulator. Idle channel noise is observed for various frequencies shown in Figure 4. Measurements for idle channel noise were made over a range of sampling frequencies.

The variation of idle channel noise (quantization noise) with sampling frequency is shown in Figure 6. At 35 kHz, the rms value of quantization noise is 100 mV. At 50 kHz, it becomes 60 mV, at 70 kHz, it becomes 40mV, at 80 kHz, it becomes 30 mV, at 90 kHz, it becomes 20 mV and finally after 95 to 100 kHz it remains constant at 10 mV or less. This is the most important and characteristic of the delta-modulation system.

Having measured the quantization noise of the delta modulator, the signal-to-noise ratio can now be evaluated. Having measured the signal power and the quantization noise at the different sampling frequencies, the signal-to-noise ratio (SNR) can be plotted as shown in Figure 7. The maximum signal-to-noise ratio of this delta modulator is 28.5 dB at 100 kHz sampling frequency. The SNR at 40 kHz is 10 dB, at 80 kHz is 21 dB. The slope of the SNR curve at 40 kHz is 9.5 dB per octave. The experimental measurements justify theoretical expression.

5.0 CONCLUSIONS

The designed delta-modulation system worked successfully. The delta-modulation system was tested for audio signals received by the standard FM Broadcast receiver. Audio output was not heard below 50 kHz sampling frequencies because of large amounts of quantization noise associated with the sampling process. As the sampling frequency was increased, quantization noise was found to decrease. Above 100 kHz sampling frequency quantization noise and circuit noise of this delta-modulation tend to remain constant at about or lower than 10 mV rms. This is the minimum noise of the system. Quantization noise, in general, obeys an inverse relationship with sampling frequency.

All delta-modulation systems have a limited dynamic range in which it accepts the input signals. Therefore, the signal-to-noise ratio (SNR) performance is limited. The measured minimum SNR of this delta-modulation system is 10 dB at 40 kHz, increasing to 15 dB at 60 kHz, 21 dB at 80 kHZ and 28.5 dB at 100 kHz. The SNR curve has a slope of 9 dB per octave at 40 kHz, therefore the theoretical performance of delta-modulation is verified.

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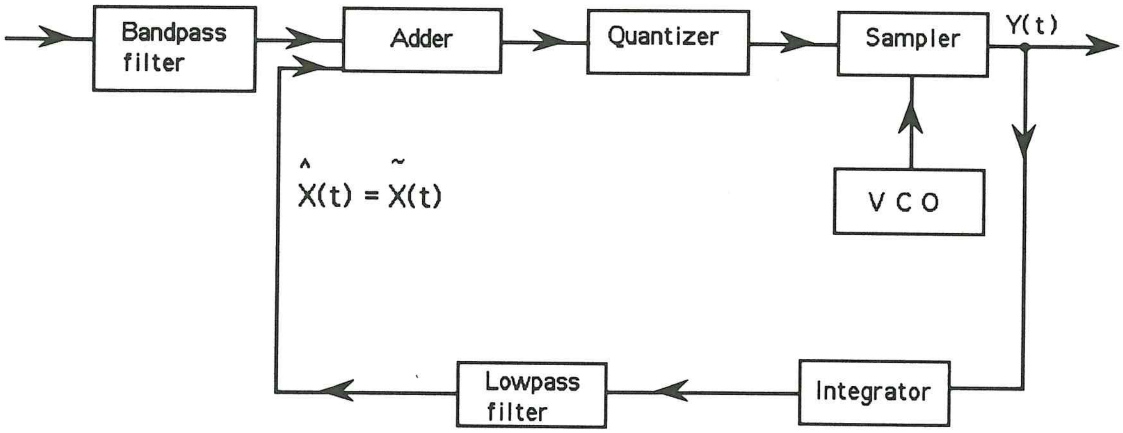
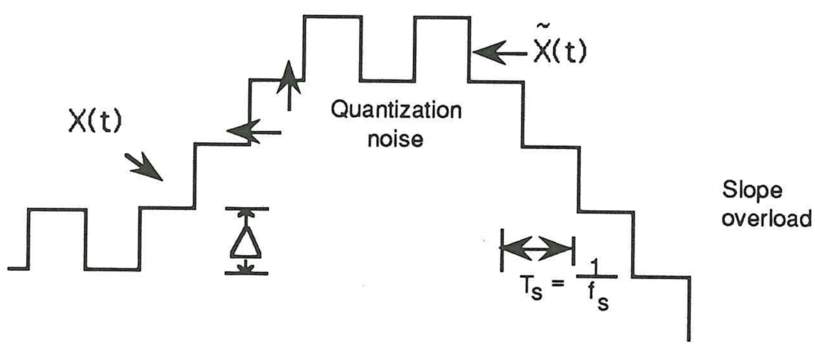


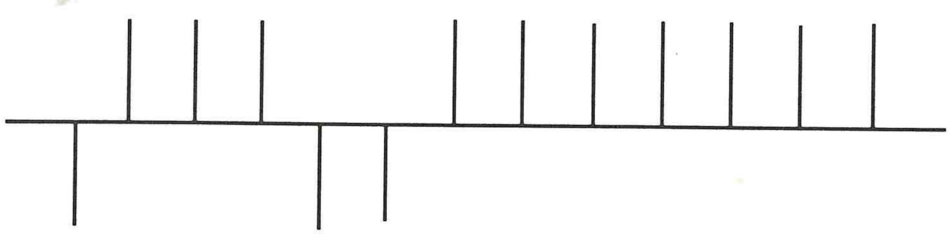
(i) Modulator

(i i) Demodulator

A. Principle of delta-modulation



B. Step-wise approximation, quantization noise, slope overload.



C. Modulator output waveform

Figure 1

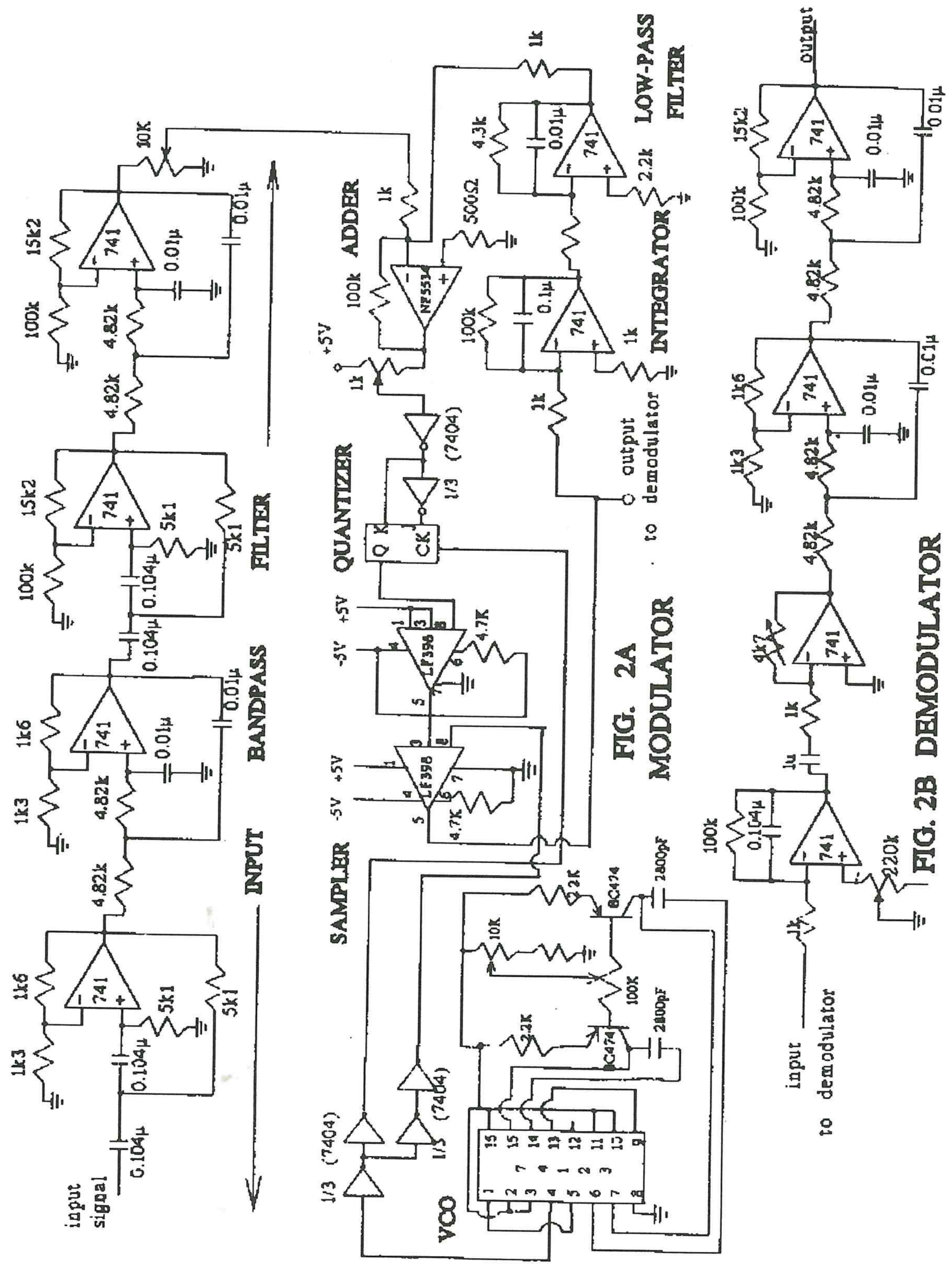


FIG. 2A MODULATOR

FIG. 2B DEMODULATOR

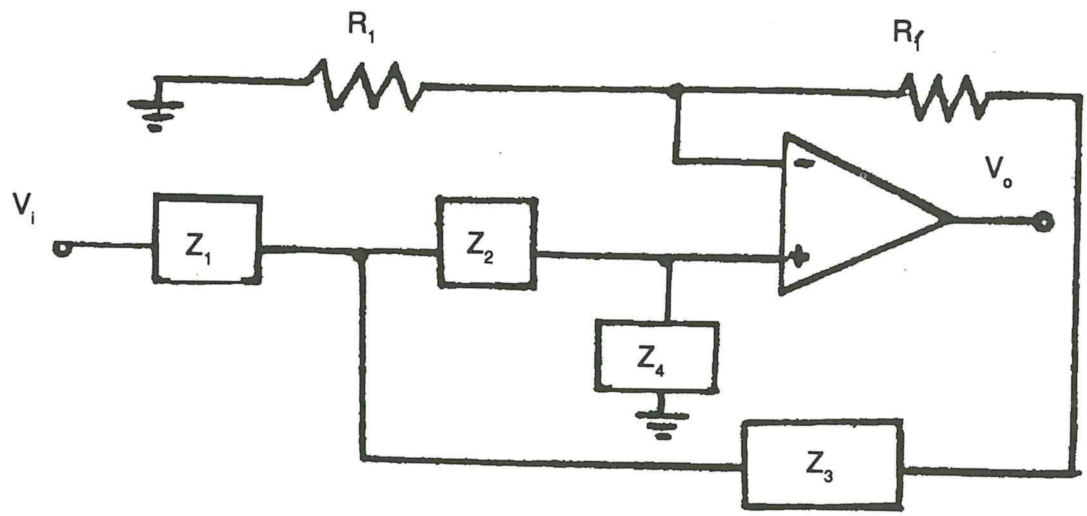
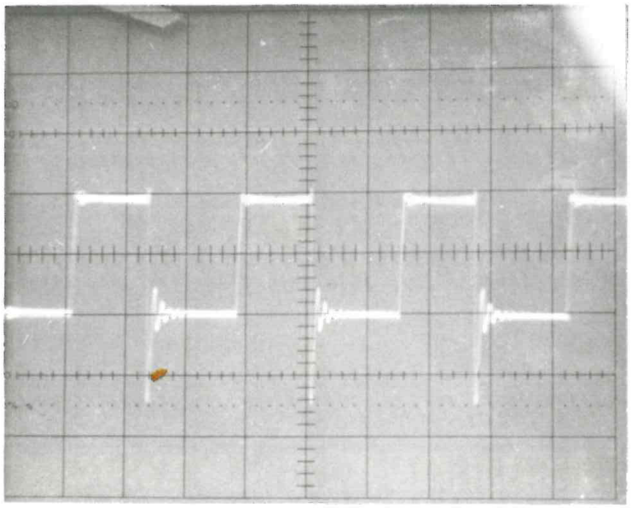
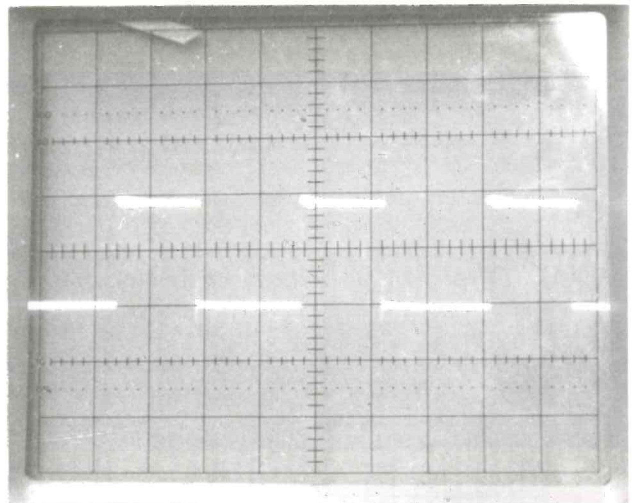


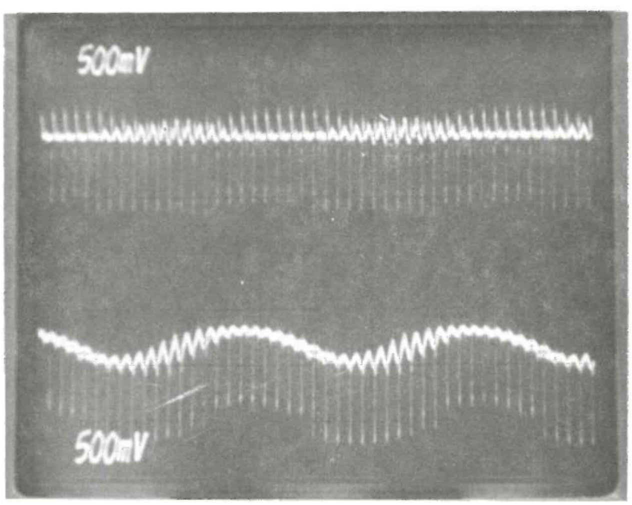
Figure 3. An active filter network.



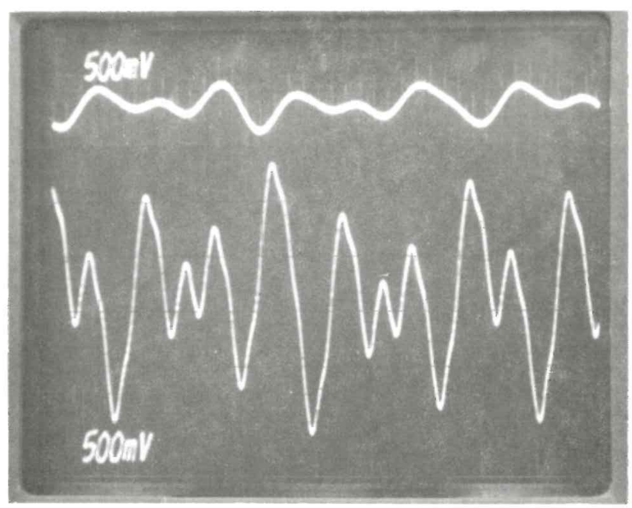
A. Voltage-controlled oscillator with spikes



B. Voltage-controlled oscillator without spikes

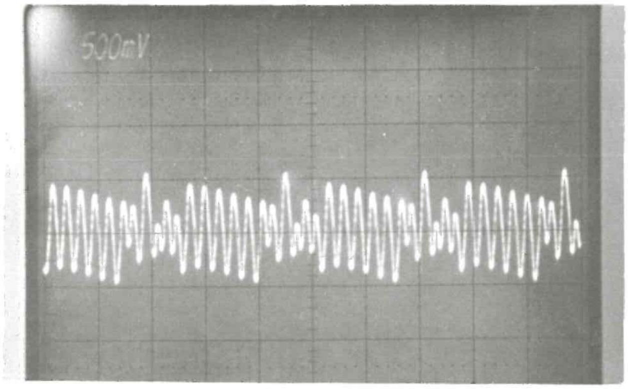


C. Idle noise output at 238 kHz

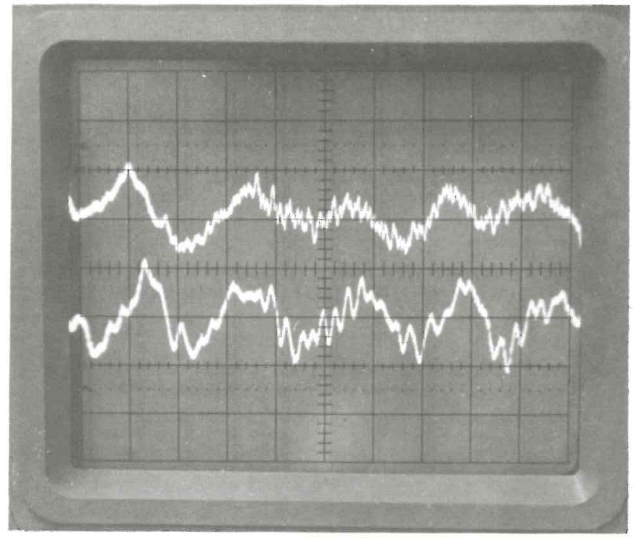


D. The demodulator and the modulator filter outputs

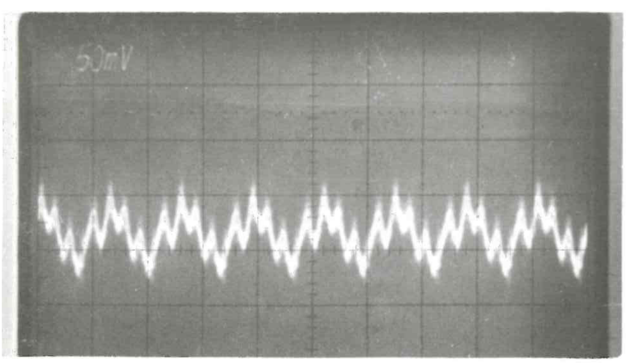
Figure 4



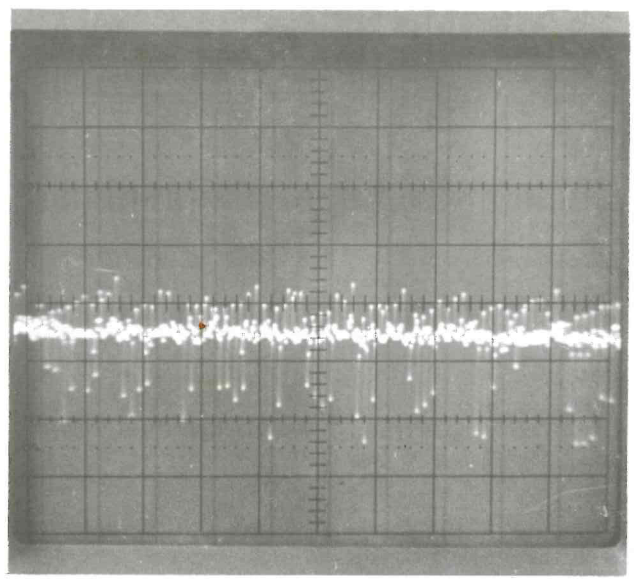
E. Noise waveform due to errors



F. Audio signal waveforms



G. Integrator output of audio signals



H. Effect of spikes on modulator output

Figure 4

N.B. Frequency scales can be multiplied by Ten for your convenience

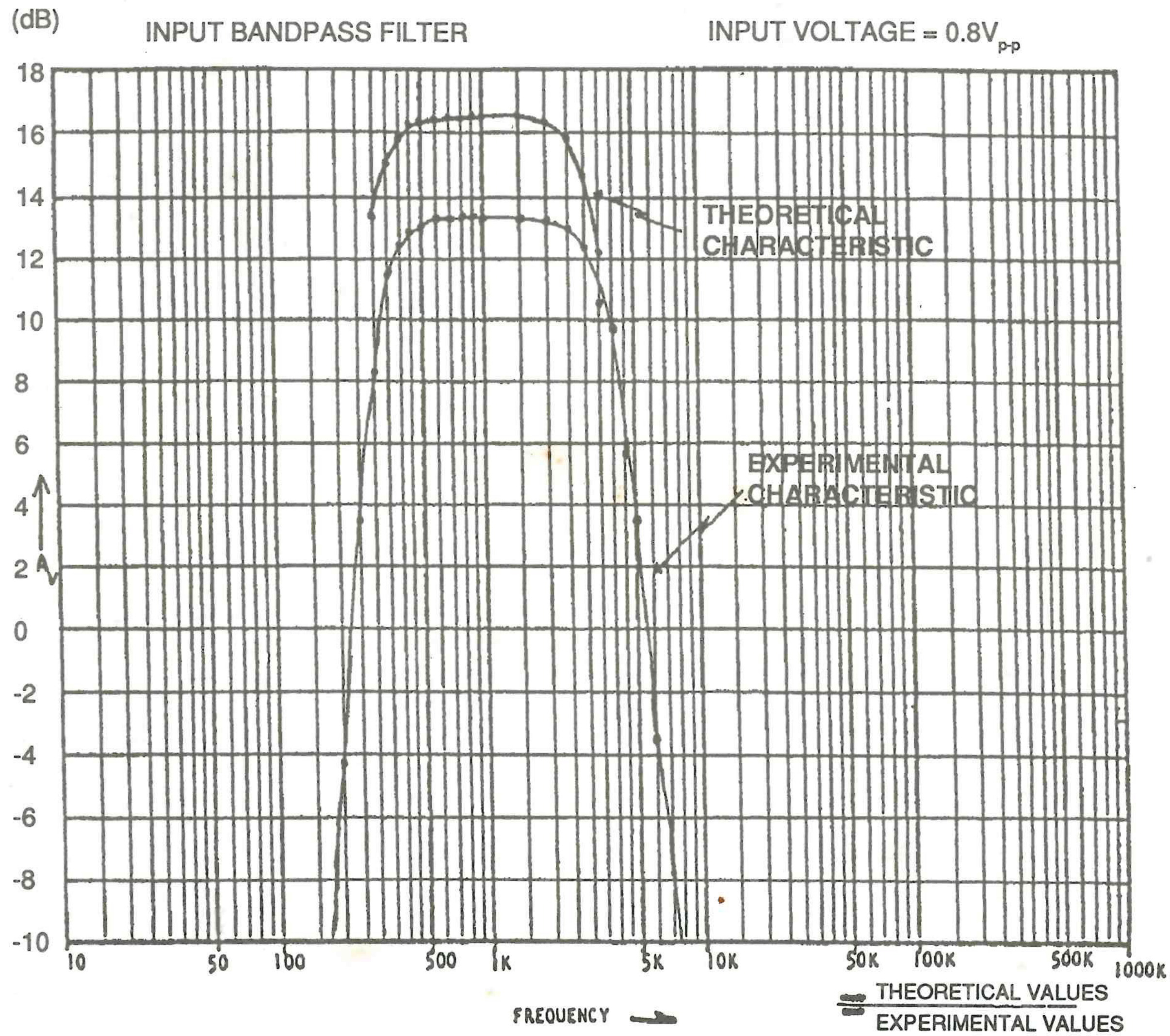


Figure 5A

N.B. Frequency scales can be multiplied by Ten for your convenience

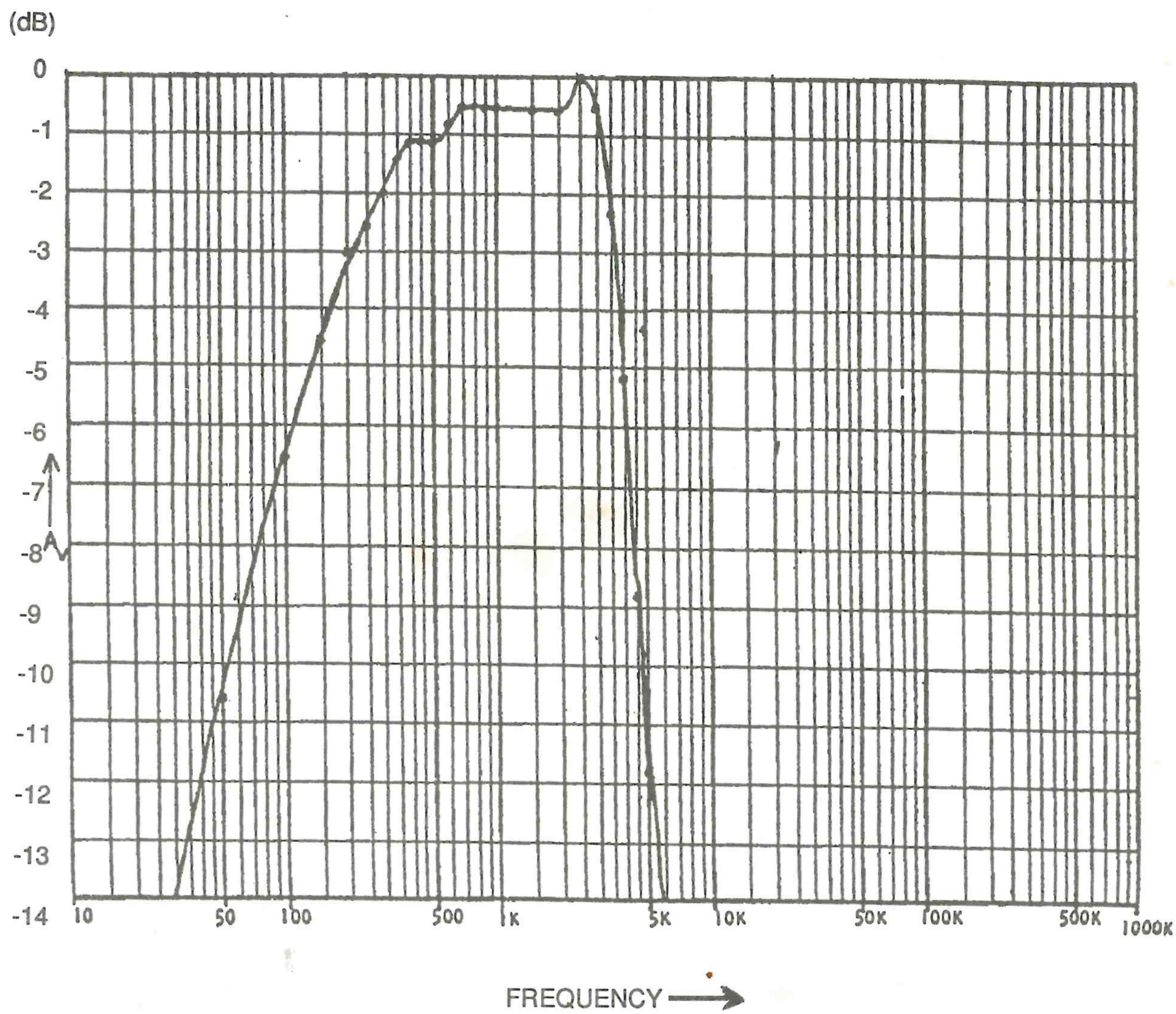


Figure 5b

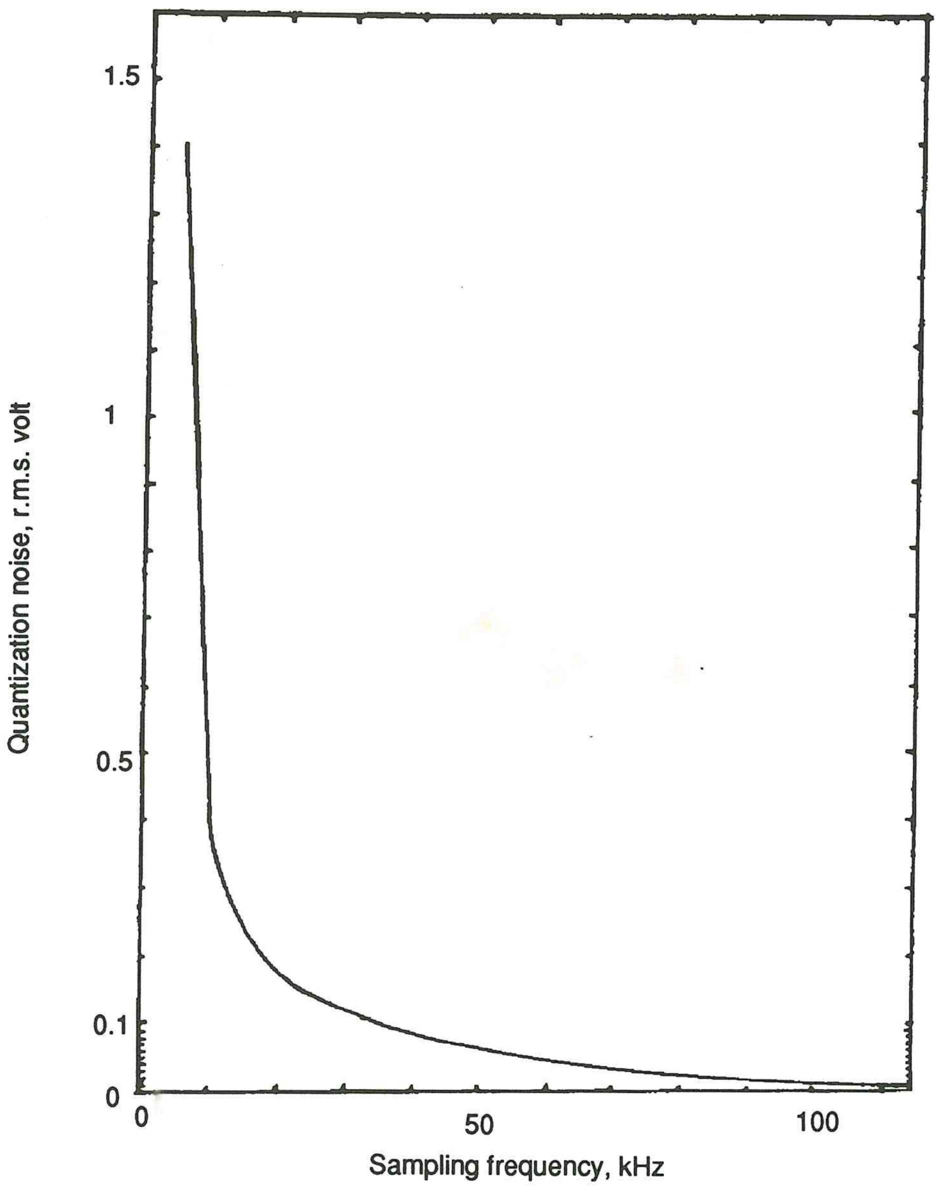


Figure 6 Quantization noise variation with sampling frequency

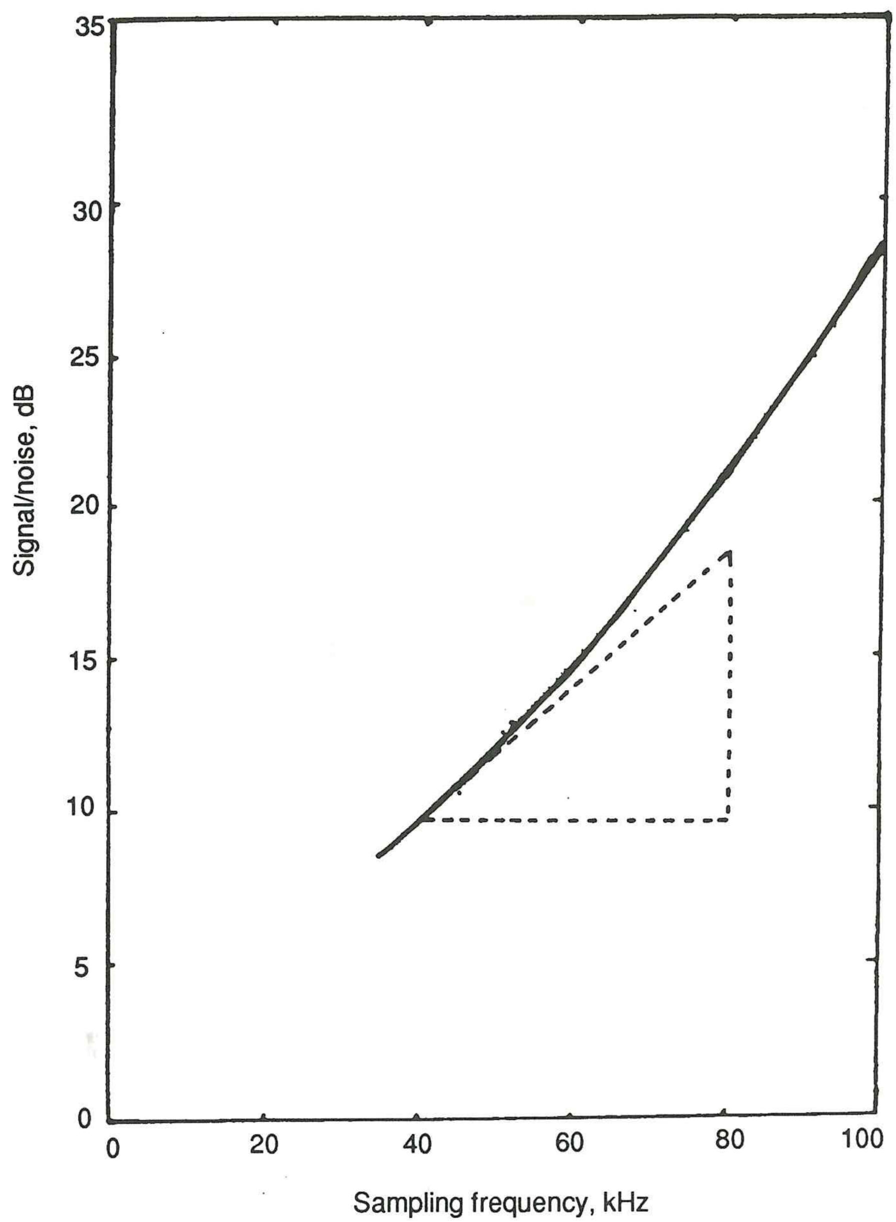


Figure 7 Signal-to noise ratio (SNR) variation with sampling frequency