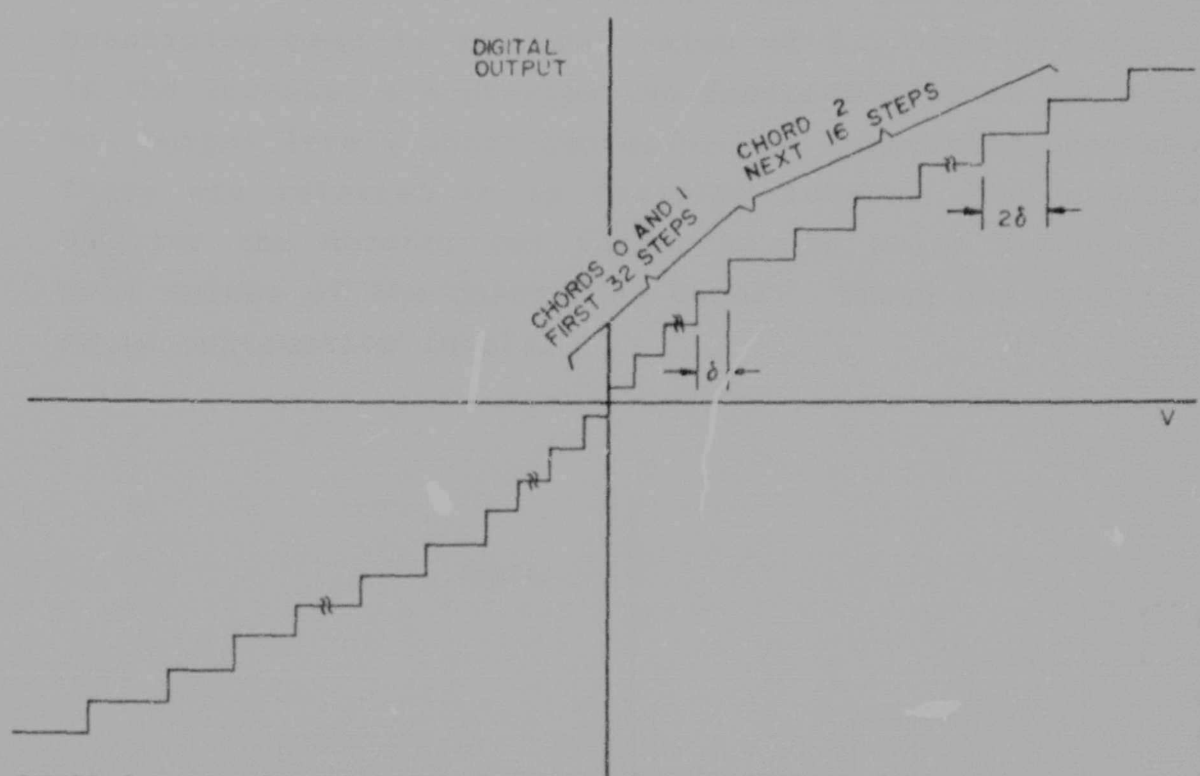

FIGURE 6 A-LAW TRANSFER FUNCTION SHOWING LINEAR STAIRCASE FUNCTION

Each chord has sixteen steps and the size of the step in each succeeding chord is double the size of the step in the preceeding chord except for the first two chords on either side of the origin. Here the steps in each chord have equal size. For the A-law function the four chords about the origin can be considered as a single segment so that the A-law characteristic is sometimes referred to as being a "13 segment" transfer function.

In order to obtain an implementation of the transfer function described above, the companding DAC is constructed in such a way that the encode and decode functions are offset by one-half step. Thus the quantizing band of the encode DAC will be centered about the decode value. This is shown in Figure 7. As an example assume that an analogue input whose amplitude lies between levels 2 and 4 is being encoded. The best quantizing code to assign to this quantizing band is its mean value of 3. Thus the DAC used in the successive approximation feedback loop of the encoder has output levels which represent the quantizing band edges. These are referred to as decision levels. The companding DAC for the decoder has output levels which represent the mean values of the quantizing bands. These are referred to as reconstruction levels.

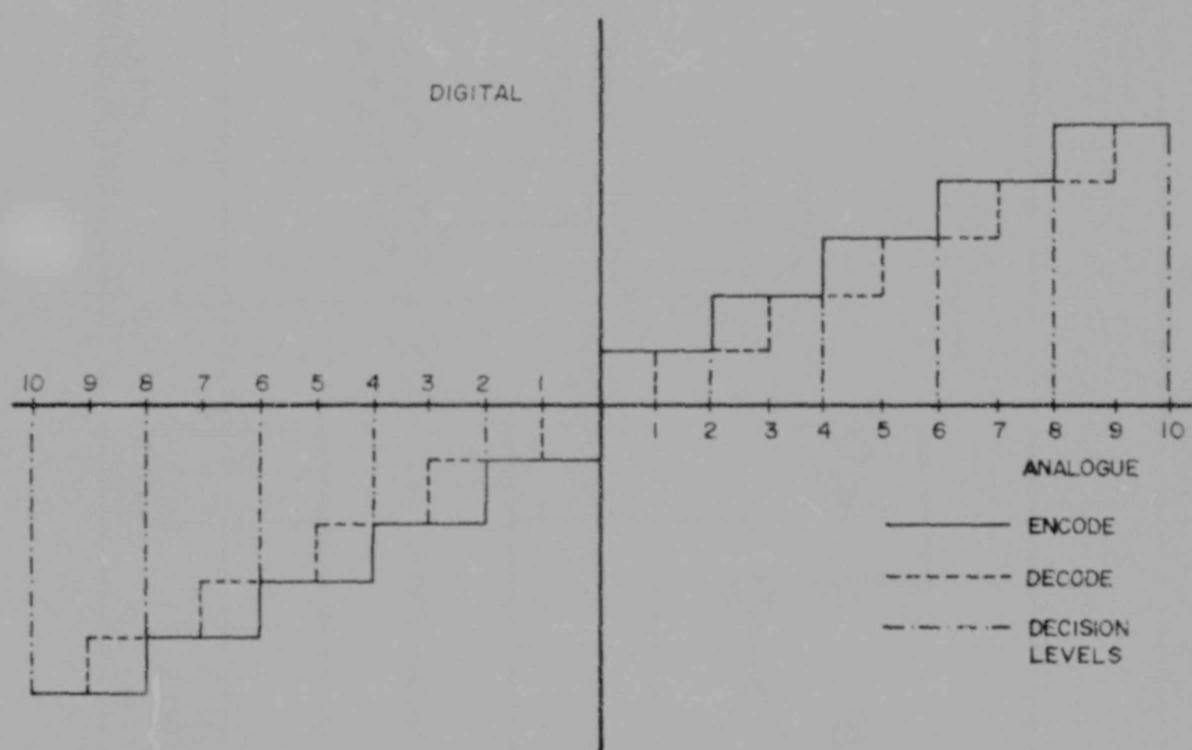

FIGURE 7 A-LAW ENCODE / DECODE CHARACTERISTIC ABOUT THE ORIGIN

In the PMI codec system a single current output DAC is used to generate outputs for either the encode or decode mode of operation. This is pin selectable as shown in figure 8. The encode mode is offset one-half step from the decode mode by means of the current generator which is switched in during the encode mode selection.

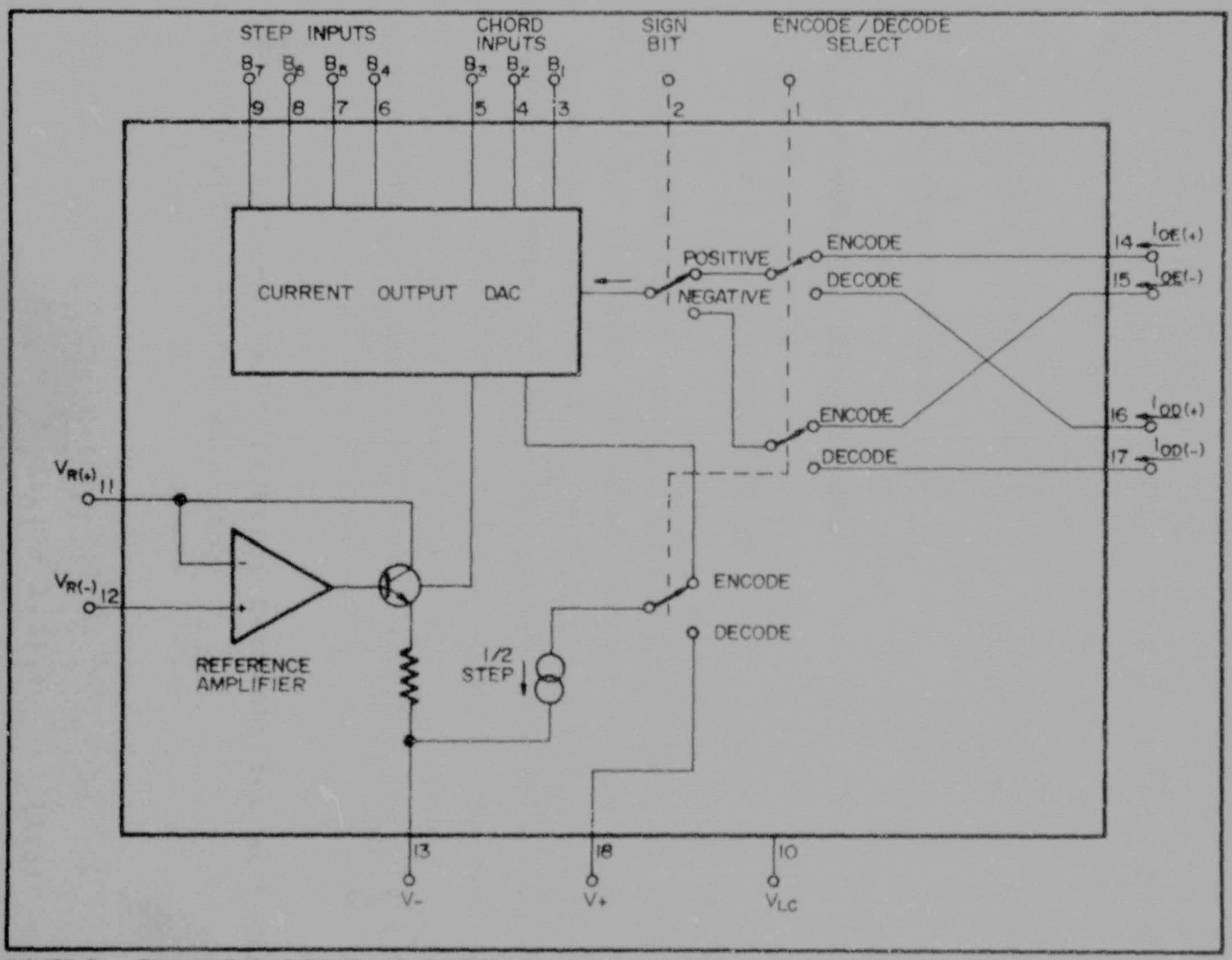

FIGURE 8   EQUIVALENT CIRCUIT OF COMPANDING DAC

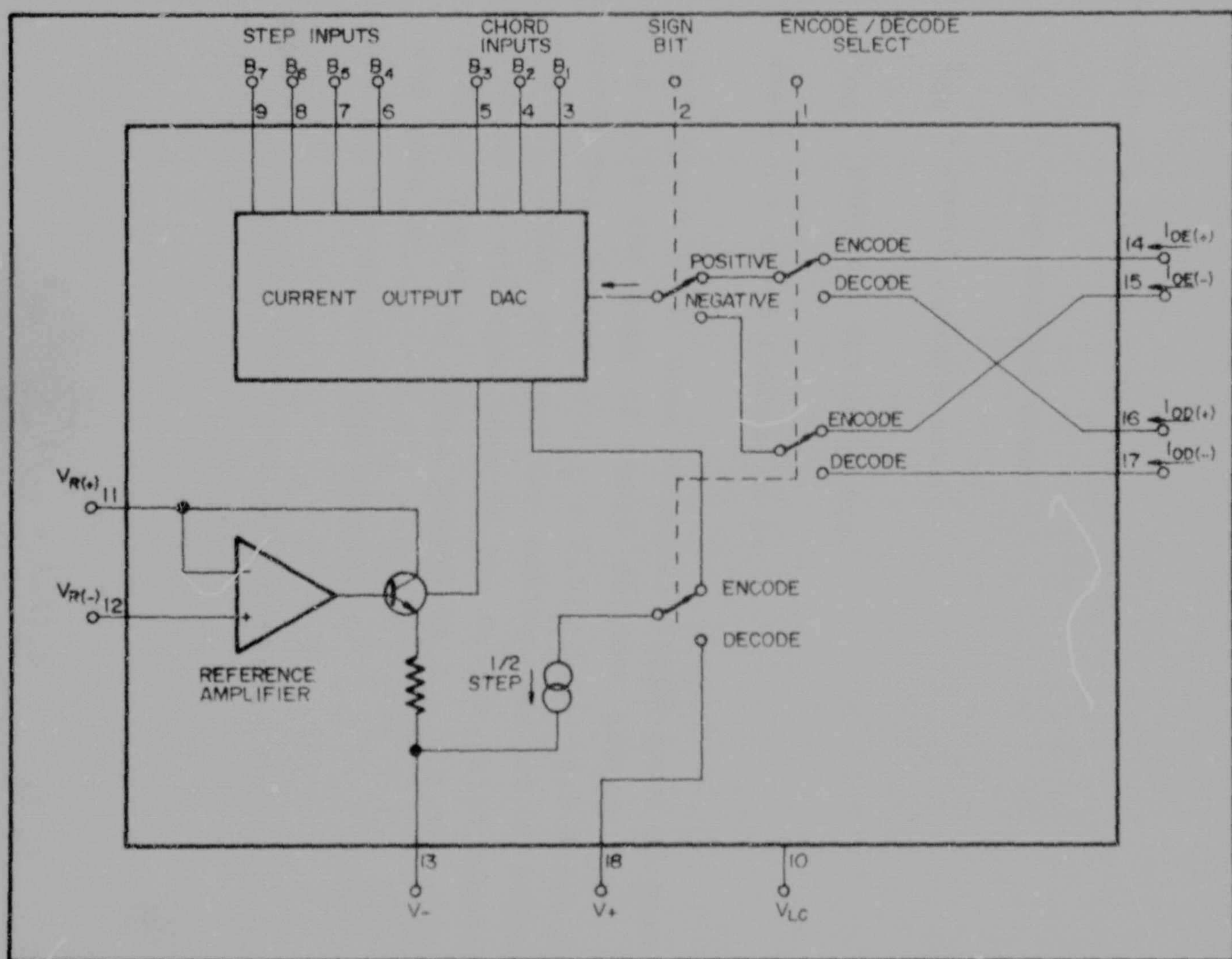

FIGURE 8 EQUIVALENT CIRCUIT OF COMPANDING DAC

Figure 9 shows the companding DAC transfer function used to implement the piece-wise linear approximation of the A-law transfer function.

The following definitions are in order :

I_N = step current size

I_{CN} = chord current size

I_{PN} = pedestal current size

where N = chord number (0 to 7)

The first chord (N=0) uses 16 equal steps each of size I_0 which is 1/16 the size of chord current source I_{C0}. The next chord, N=1, begins at $I_{C0} + 1.5I_0$ i.e. 16.5 steps from the origin. This is done by making use of a pedestal current which is directed towards the output, with magnitude $I_{C0} + 1.5I_0$. The third chord begins at $I_{C0} + 1.5I_0 + I_{C1} + 1.5I_1$ and ends at $I_{C0} + 1.5I_0 + I_{C1} + 1.5I_1 + I_{C2} + 1.5I_2$ and so on for the rest of the chords.

The process continues with pedestal currents for each chord number N described by the equation:

$$I_{PN} = \sum_{i=0}^{N-1} (I_{Ci} + 1.5I_i) \qquad (1.3)$$

with $I_{P0} = 0$

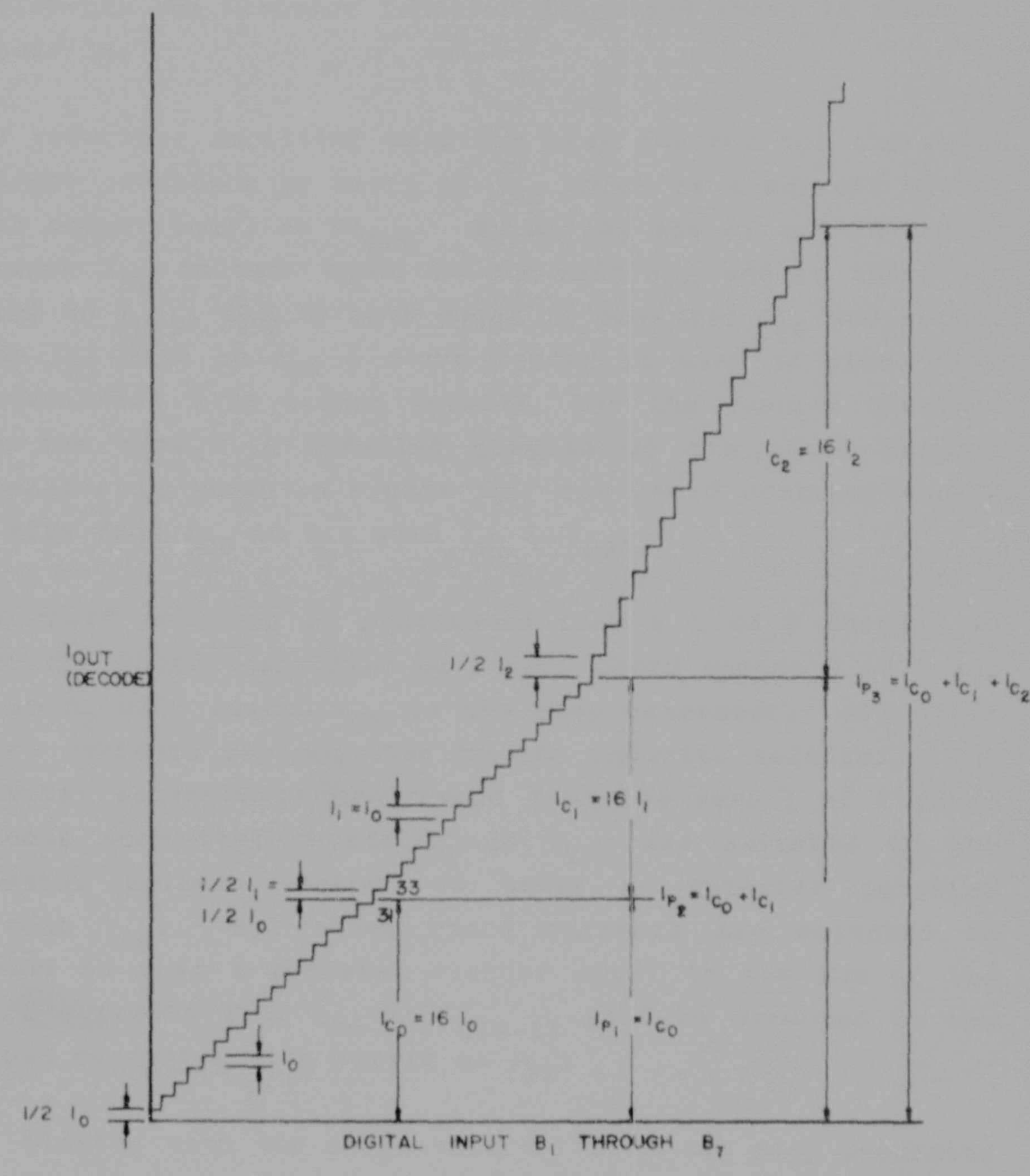

FIGURE 9 CONSTRUCTION OF THE COMPANDING DAC TRANSFER FUNCTION

The functional diagram of the companding DAC which implements the transfer function discussed above is shown in Figure 10.

The reference amplifier sets the bias current for the chord current generator by means of I_{C7} which is a current mirror with output equal to $2I_{REF}$. By making use of a R-2R ladder network I_{C6} is made equal to one-half I_{C7} and is therefore equal to I_{REF}. I_{C5} is made equal to one-half I_{C6} and so on. From I_{C3} down to I_{C0} a slave ladder is used in place of a conventional R-2R ladder network, but the results obtained are the same. A detailed diagram of the chord current generator is shown in Figure 11. The chord currents double in size from I_{C0} to I_{C7} with $I_{C1} = I_{C0}$.

The chord selector is programmed from a 1 of 8 decoder so that the chord identified by binary chord number N on leads B_1 to B_3 will switch I_{CN} to the step generator. All other chord currents are switched to the pedestal selector. The pedestal selector is programmed from the same 1 of 8 chord decoder such that chords I_0 to I_{N-1} are switched to the pedestal selector output in order to generate pedestal current I_{PN}. All other chord currents are switched to ground so that a pedestal current equal to the sum of the chord currents from I_{C0} to $I_{C(N-1)}$ will be directed to the output current switch matrix as I_{PN}.

I_{CN} flowing into the chord selector from the step generator is equal to 16 step currents, where a step current is equal to the current step caused by changing the least significant bit in the chord of interest. A detailed diagram of the step current generator is shown in figure 12. The step current generator has the ability to introduce a current I_E into the output so as to provide the one-half step offset required when the system is operating in the encode mode. This one-half step offset current is controlled by the E/D pin.

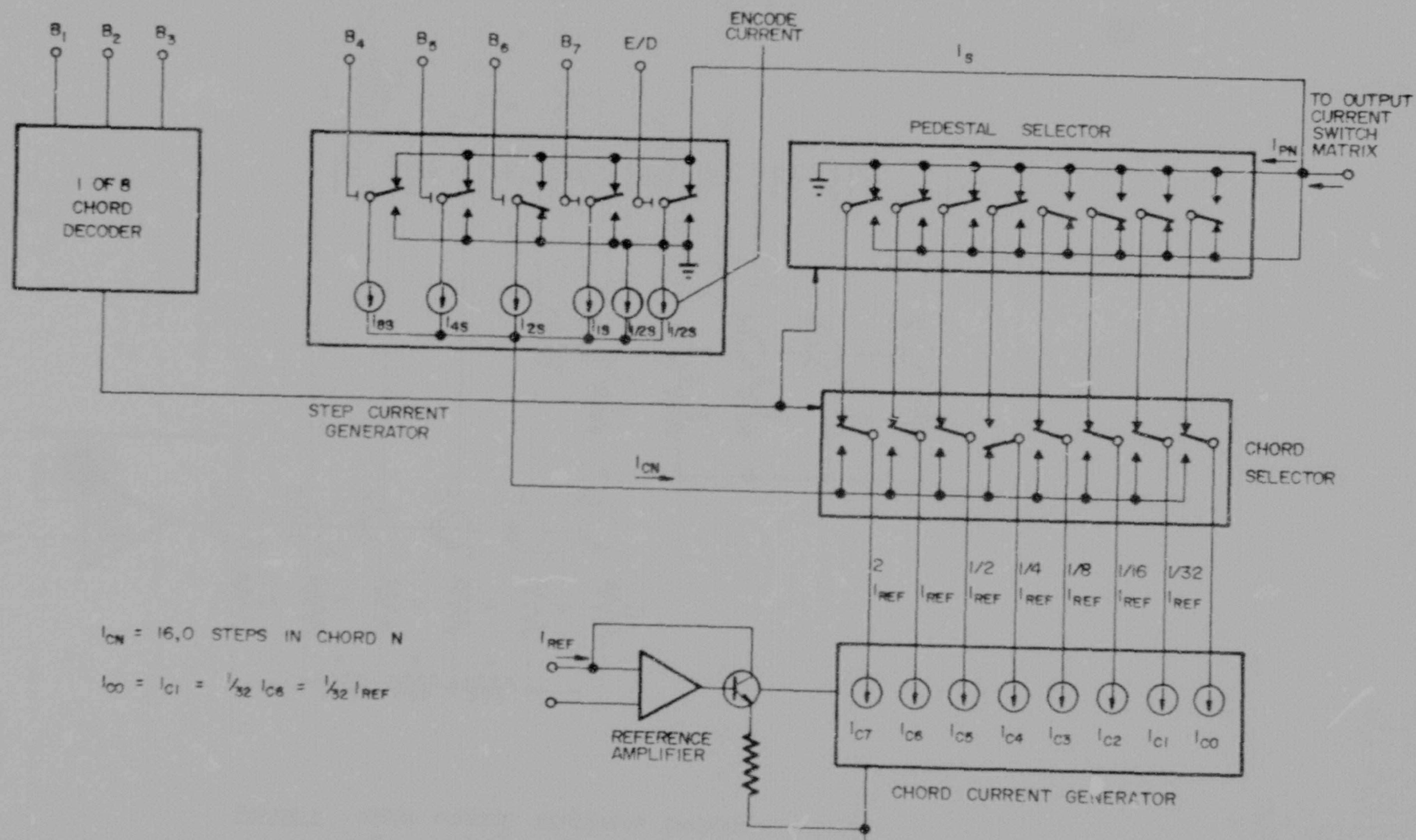

FIGURE 10. COMPANDING DAC FUNCTIONAL DIAGRAM

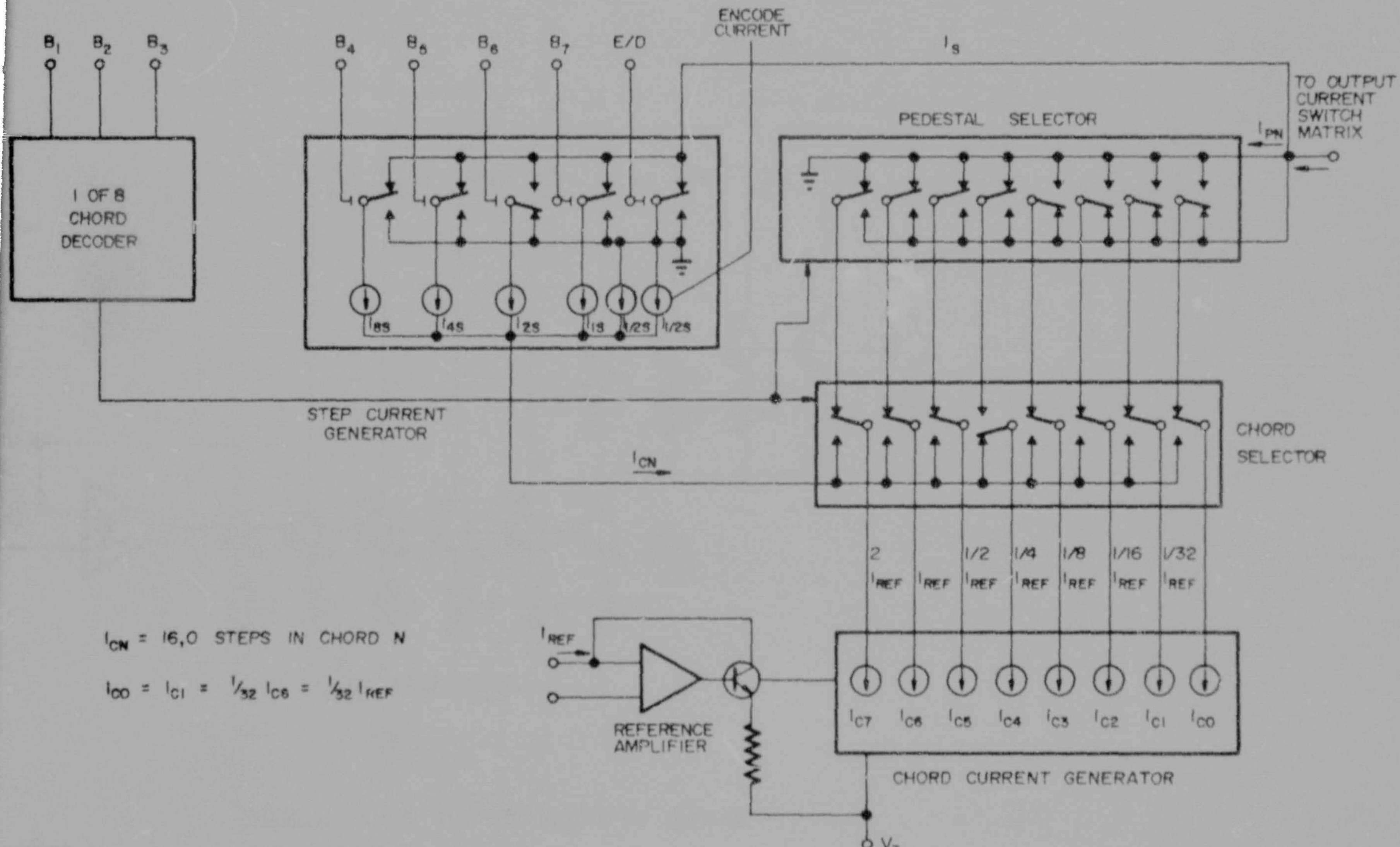

FIGURE 10. COMPANDING DAC FUNCTIONAL DIAGRAM

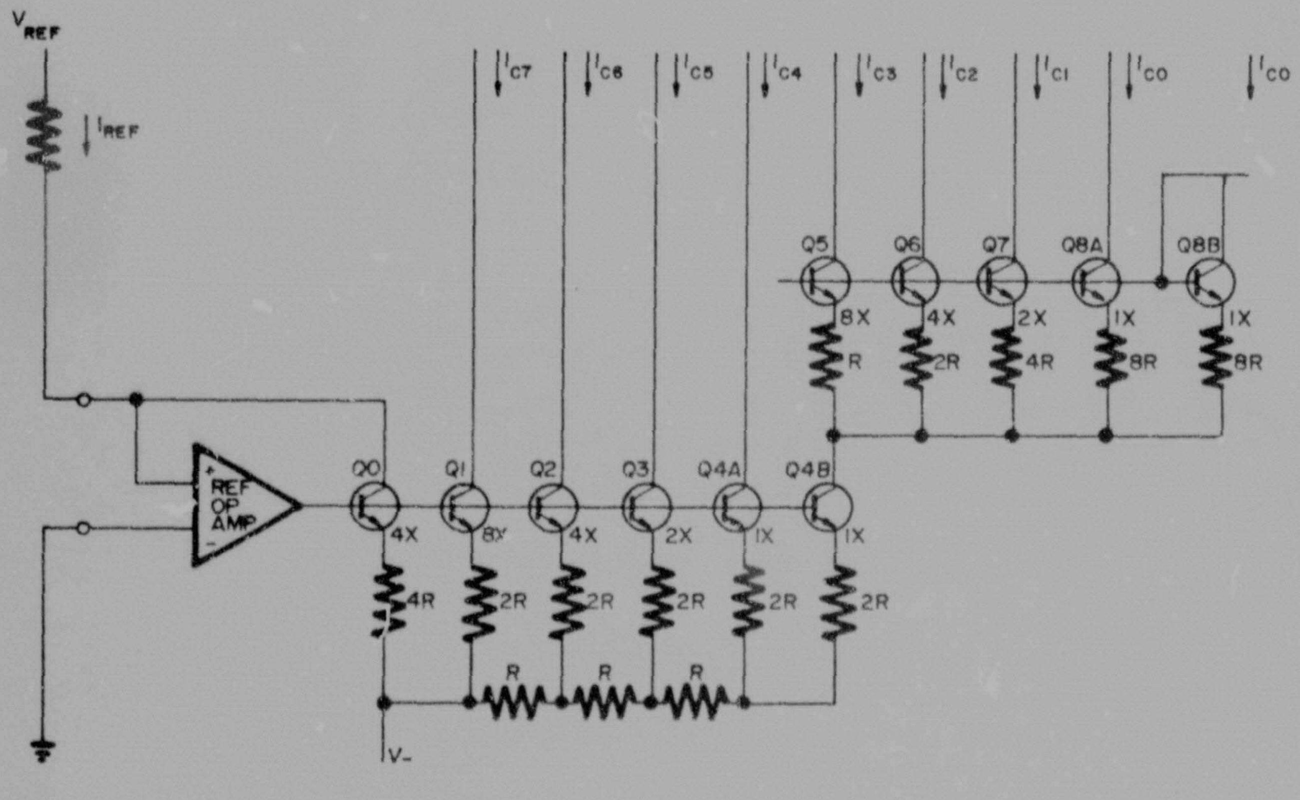

FIGURE 11 CHORD CURRENT GENERATOR DIAGRAM.

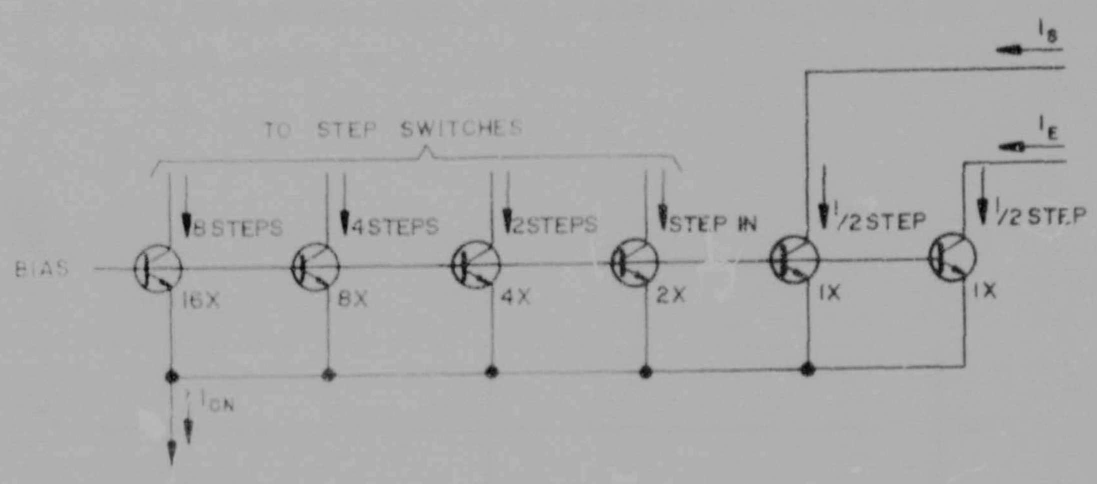

FIGURE 12 STEP CURRENT GENERATOR

1.3.3 Quantization distortion

The theory of quantization distortion or noise is covered exhaustively in the literature [Cattermole, 1969; Betts, 1970; Taub and Schilling, 1971] and the intention here is to give a brief summary.

Quantization distortion is very diffe ent in nature from the familiar harmonic and intermodulation distortion present in analogue communication systems. This is due to the fact that the transfer characteristic used, for encoding and decoding, is a discontinuous function, consisting of a number of discrete steps and thus any analogue waveform can only be approximated. The error due to this approximation manifests itself as noise.

Consider a linear codec, a section of which is shown in figure 13. Let the instantaneous amplitude of the analogue signal be v and the nearest quantizing level be v_i. $(\Delta v)_i$

defines the step over which sampled values will be assigned to quantum level v_i.

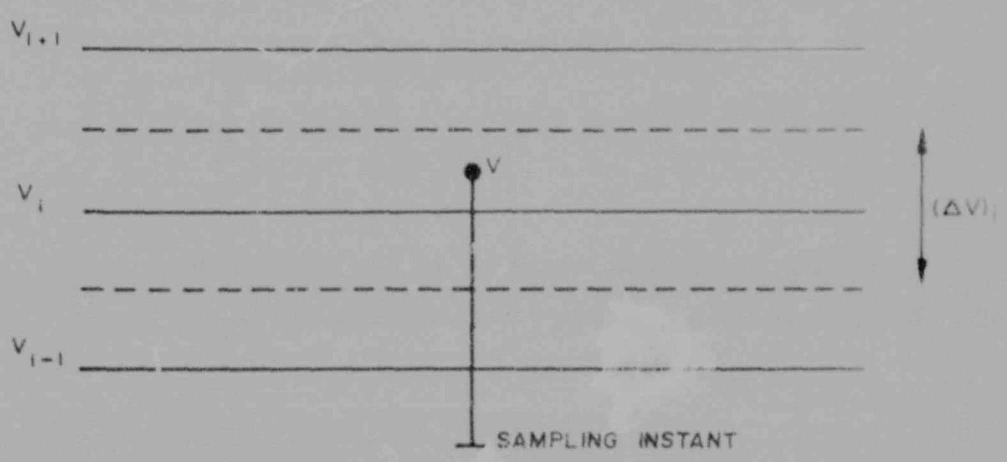

FIGURE 13 QUANTIZATION OF [illegible] ANALOGUE SIGNAL

The quantizing level v_i is implied as being a decoder level. The levels $v_i-(\Delta v)_i/2$ and $v_i+(\Delta v)_i/2$ are implied as being encoder levels.

The quantizing error is given by

$$v - v_i \tag{1.4}$$

and the mean square error voltage is

$$\overline{(v - v_i)^2} \tag{1.5}$$

All the signal amplitudes within the range $v_i-(\Delta v)_i/2$ to $v_i+(\Delta v)_i/2$ will be referred to the i^{th} level and the mean square error voltage associated with this level is σ_i^2 given by

$$\sigma_i^2 = \int_{v_i - (\Delta v)_i/2}^{v_i + (\Delta v)_i/2} (v - v_i)^2 \, p(v) \, dv \qquad (1.6)$$

where p(v) is the probability of an instantaneous voltage v falling within the range $v_i-(\Delta v)_i/2$ to $v_i+(\Delta v)_i/2$

If the step size is small we can assume that the signal is uniformly distributed over the step and therefore $p(v)=p(v_i)$.

Thus the above equation reduces to

$$\sigma_i^2 = \frac{(\Delta v)_i^3 \, p(v_i)}{12} \qquad (1.7)$$

The total mean square quantizing noise, σ^2, is given by

$$\sigma^2 = \sum_i \sigma_i^2 = 1/12 \sum_i (\Delta v)_i^3 \, p(v_i) \qquad (1.8)$$

$$= 1/12 \sum_i (\Delta v)_i^2 \, p(v_i) \, (\Delta v)_i \qquad (1.9)$$

All the signal amplitudes within the range $v_i-(\Delta v)_i/2$ to $v_i+(\Delta v)_i/2$ will be referred to the i^{th} level and the mean square error voltage associated with this level is σ_i^2 given by

$$\sigma_i^2 = \int_{v_i - (\Delta v)_i/2}^{v_i + (\Delta v)_i/2} (v - v_i)^2 \, p(v) \, dv \qquad (1.6)$$

where p(v) is the probability of an instantaneous voltage v falling within the range $v_i-(\Delta v)_i/2$ to $v_i+(\Delta v)_i/2$

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$$= 1/12 \sum_i (\Delta v)_i^2 \, p(v_i) \, (\Delta v)_i \qquad (1.9)$$

The probability of the signal amplitude being within the i^{th} step is

$$P_i = \int_{v_i - (\Delta v)_i/2}^{v_i + (\Delta v)_i/2} p(v)\, dv \qquad (1.10)$$

and since $(\Delta v)_i$ is small compared with the total amplitude range

$$P_i \approx p(v_i)\,(\Delta v)_i \qquad (1.11)$$

Thus

$$\sigma^2 \approx 1/12 \sum_i (\Delta v)_i^2\, P_i \qquad (1.12)$$

As we have linear quantization i.e. all steps are of equal size

$$\sum (\Delta v)_i^2\, P_i = (\Delta v)^2 \qquad (1.13)$$

Therefore

$$\sigma_{lin}^2 = (\Delta v)^2/12 \qquad (1.14)$$

This is the fundamental equation for the mean square quantizing noise introduced by a linear quantizing law.

As discussed earlier, the piece wise linear approximation to the A-law is comprised of a number of segments, the steps within each segment being linear. Thus the piece-wise linear approximation to the A-law is made up of a number of linear segments and the mean square quantizing noise for each chord or segment is given by the above equation. This is substantiated by Pearce [1982]. Here, he describes a three chord codec used for FDM hypergroup translation, and gives the rms quantizing noise as:

$$N_q = \frac{2V}{2^n \quad 12}\left\{\int_{\emptyset}^{V/4} p(v)\,dv + 4\int_{V/4}^{V/2} p(v)\,dv + 16\int_{V/2}^{V} p(v)\,dv\right\}^{1/2}$$

where n = number of bits

V = peak input range of the codec

and V/4, V/2, V are the chord end-points

(1.15)

In addition, Fourier analysis can be used to calculate the signal to quantization distortion produced by the quantizing process and this is covered by Tsividis and Zee [1979].

Tsividis and Zee consider a sinusoidal signal x(t) being applied to the input of a step quantizer with transfer characteristic given by y=y(x). The function y=y(x) gives the decoder output level (filtered or unfiltered) in terms of a continuous amplitude x. The input signal periodic output can be represented by a Fourier series:

$$y(t) = b_{\emptyset} + \sum_{n=1}^{\infty} b_n \cos(\omega t + \theta_n) \qquad (1.16)$$

The dc term, $b_{\emptyset}$, is of no consequence in the present work.

From (1.16) the signal to quantization distortion ratio is given by:

$$\frac{S}{D_q} = \frac{b_1^{\,2}}{\sum_{n=2}^{\infty} b_n^{\,2}} \qquad (1.17)$$

1.3.4 Gain tracking

Gain tracking refers to the ability of a system to linearly track its input power level. In a system displaying an ideal gain tracking characteristic, any change in the input level to the system must be matched exactly by the same change in the output level. Gain tracking is sometimes referred to in the literature as gain deviation.

Considering the method of Fourier analysis employed in the previous section, and letting the amplitude of the sinusoidal input x(t) be A, the quantizer gain is given by the following equation:

$$G = \frac{b_1}{A} \qquad (1.18)$$

1.4 Definitions

It is appropriate at this stage to define certain terms which are used extensively throughout the present work. The terminology is concerned mainly with the principle of half channel measurements.

Half channel measurement for encoder

The half channel performance (gain tracking, quantization distortion) for an encoder is equal to the overall performance of the encoder under test and a decoder which has an ideal performance for the waveform used to stimulate the encoder. This is shown in schematic form in figure 14.

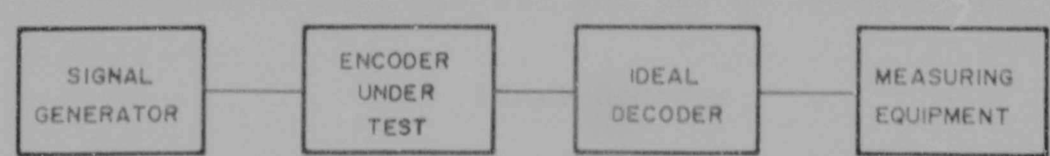

FIGURE 14 HALF CHANNEL MEASUREMENT FOR ENCODER

Half channel measurement for decoder

The half channel performance of a decoder is equal to the overall performance of an ideal encoder feeding the decoder under test when measured from end to end with a specified waveform. This is shown in schematic form in figure 15.

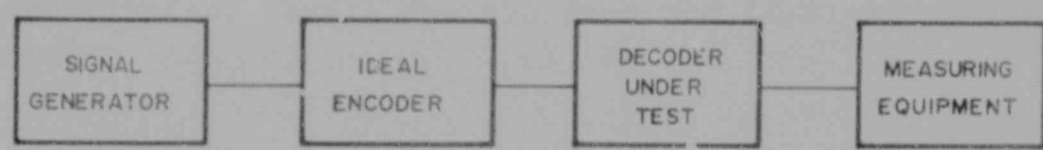

FIGURE 15 HALF CHANNEL MEASUREMENT FOR DECODER

Reference curve

The reference curve for a particular parameter (gain tracking, quantization distortion) is the curve that would be obtained with an ideal encoder and ideal decoder connected in cascade.

Ideal encoder

An ideal encoder is one in which the decision levels conform exactly to the decision levels defined in CCITT recommendation G711 [1976a] and has no impairments.

Ideal decoder

An ideal decoder is one in which the reconstruction levels conform exactly to the reconstruction levels defined in CCITT recommendation G711 [1976a] and has no impairments.

Imperfect encoder

A imperfect encoder is defined as one in which the decision levels deviate from the ideal encoding law as specified by CCITT.

Imperfect decoder

A imperfect decoder is defined as one in which the reconstruction levels deviate from the ideal as specified by CCITT.

Unfiltered encoder

An unfiltered encoder is defined as an encoder which is not preceeded by a normal anti-aliasing or send filter, but is fed with a test signal with spectrum which is limited by some means to ensure that aliasing of energy is acceptably low. The test signal may be continuous in time as the encoder has a on-chip sample and hold.

Unfiltered decoder

An unfiltered decoder is defined as a decoder which has no reconstruction filter associated with it in the normal sense, but which produces a output which is essentially PAM with pulse height equal to the decoder output level, which is held constant during the intervals between samples, apart from a short time during which it slews from the previous output to the current value.

1.5 Literature survey of methods proposed for measuring the performance of codecs.

Until recently PCM systems were implemented in an analogue to analogue configuration in order to provide increased capacity on existing cable junction links between electro-mechanical exchanges. Consequently, the measurement methods used to determine the performance of the systems required signal generation and signal measurement in the analogue domain. With the introduction of electronic switching the encoding and decoding functions became separated by means of a switching matrix introduced between them. As a result seperated performance measurements are required. For the encoder this entails signal generation in the analogue domain and signal measurement in the digital domain. For the decoder signal generation in the digital domain and signal measurement in the analogue domain is required.

For the measurement of quantization distortion and gain tracking two basic methods are proposed, viz. the frequency domain subtraction method and the time domain subtraction method. The frequency domain subtraction method is used in all commercially available instruments. The time domain subtraction method is proposed in the literature but, as far as the author is aware, has not been used in any commercial instrument.

The time domain subtraction method involves comparing the input signal to the device under test with the output signal from the device. Kikkert [1971] describes a method whereby it is possible to measure the quantization distortion of a PCM system (full channel or half channel) by using the arrangement shown in figure 16. W is defined as the difference between the input signal to the system(V) and the output of the system (U), and is termed the error due to the quantizing process. Thus the signal to quantizing distortion ratio would be given by Q=V/W.

If separated tests are to be conducted on the encoder and decoder an analogue to digital and digital to analogue converter would be required. For the encoder test the input signal(V) would be in the analogue domain and the output signal (U) would be in the digital domain requiring a digital to analogue converter before the comparison with the input signal can take place. For the decoder test the input signal would need to be in the digital domain thus requiring a analogue to digital converter or reference encoder. The above is required assuming that signal generation and signal measurement are done in the analogue domain.

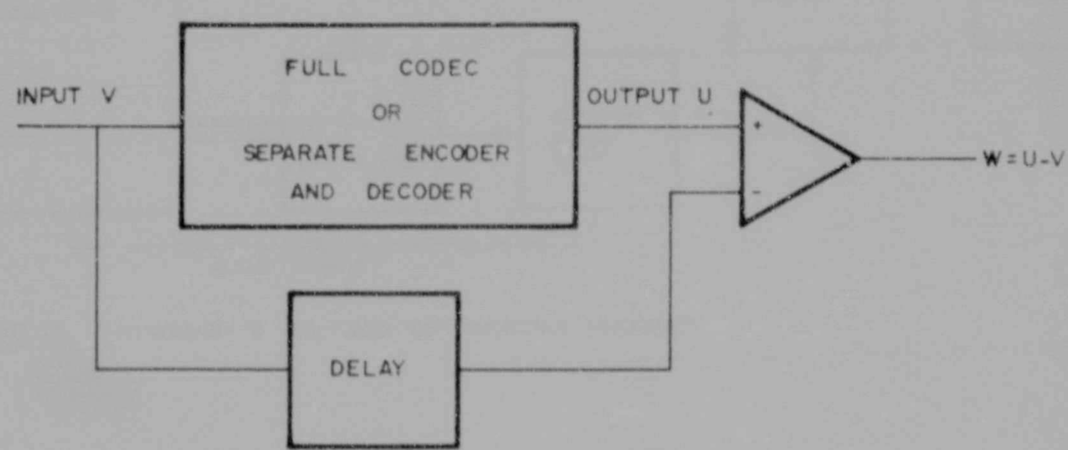

FIGURE 16 TIME DOMAIN SUBTRACTION METHOD OF MEASURING QUANTIZATION NOISE

Stoddart [1971] proposes a method whereby it is possible to measure the quantizing error introduced in the encoder. This is shown in figure 17. The signal generator produces a 2 kHz analogue sinusoidal signal and a 8 bit PCM word which corresponds to the 2 kHz signal being encoded by an ideal encoder for different amplitudes. The output of the encoder is then compared with the 8 bit PCM word and the difference is used to trigger the error counter.

A sinusoidal signal of frequency 2 kHz is chosen so that it is possible (using an 8 kHz sampling frequency) to sample the analogue signal at its peak and zero amplitude points. A similar circuit is used for testing the decoder.

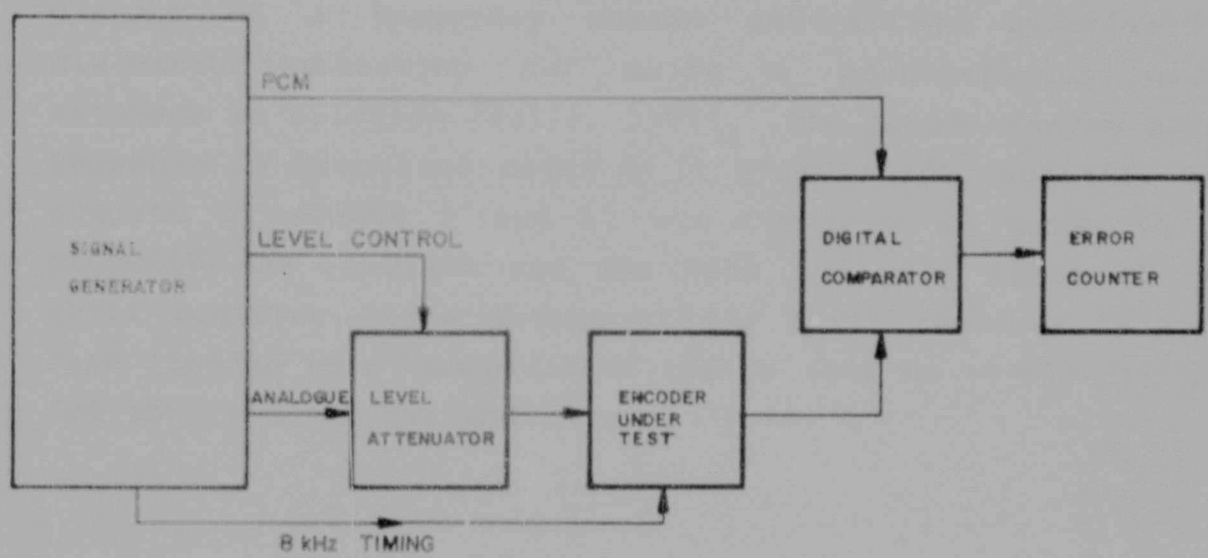

FIGURE 17 STODDART'S METHOD OF TESTING ENCODER

The above two methods, specifically that proposed by Stoddart, use a static level method for testing the codec. This involves testing at each of the codec's 256 quantizing levels, i.e. both positive and negative, to ensure that they function according to the transfer characteristic discussed earlier.

The static level method has two major drawbacks [Attridge et al., 1979]. Firstly, the test results do not directly relate to the CCITT specifications. Secondly, the detection of dynamic and frequency dependent errors is not possible when using this method. An example of a dynamic error is the effect of the sample and hold circuit having to slew between opposite ends of the amplitude range, giving a larger error than when two successive samples are close to each other. Consequently, time domain subtraction is not a good method for determining the performance of integrated circuit codec's.

As mentioned previously, the second method proposed is that of frequency domain subtraction. Figure 18 gives the block diagram of a frequency domain subtraction quantization distortion measuring set using a pseudo-random noise sequence as stimulus [Rolls, 1980]. The pseudo-random noise sequence is generated using a 17 stage shift register. The outputs of stages 3 and 17 are combined by means of an EXCLUSIVE OR function and fed back into the input of the shift register. The output of the shift register is then band limited to a bandwidth of 100 Hz (450 Hz to 550 Hz) and fed into the encoder under test.

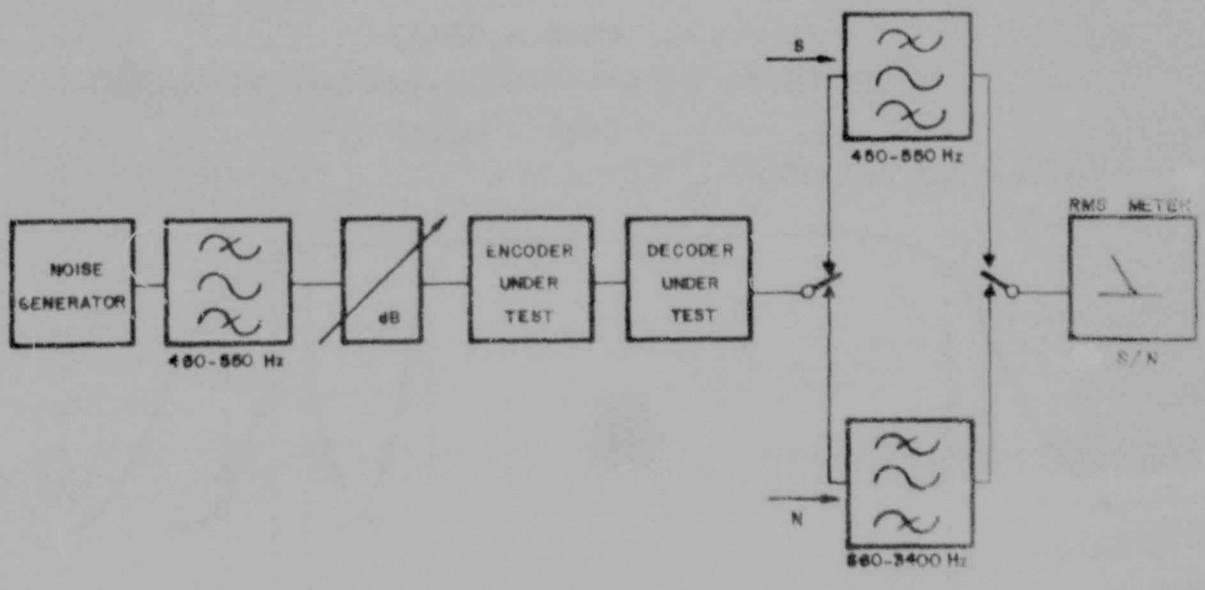

FIGURE 18 FREQUENCY DOMAIN SUBTRACTION METHOD FOR MEASURING QUANTIZATION NOISE

The output signal from the decoder is split into two paths. In the first path the signal passes through a 450-550Hz bandpass filter, the purpose of which is to filter out the rest of the channel frequencies, leaving the noise stimulus signal alone. This is measured by the rms meter to give a value S. In the second path the signal passes through a 860-3400 Hz bandpass filter, which filters out the stimulus signal leaving the quantization distortion components. This is measured by the rms meter giving a value N.

The ratio S/N is then taken giving the signal to quantization distortion ratio.

The frequency domain subtraction method for measuring quantization distortion is shown in figure 19. As the quantization distortion measured is for the range 860-3400Hz, a correction factor is used to quote the quantization distortion over the full channel spectrum viz. 300-3400Hz. This correction factor is given by:

$$10 \log \frac{(3400 - 860)}{(3400 - 300)} = -0.87 \text{ dB}$$

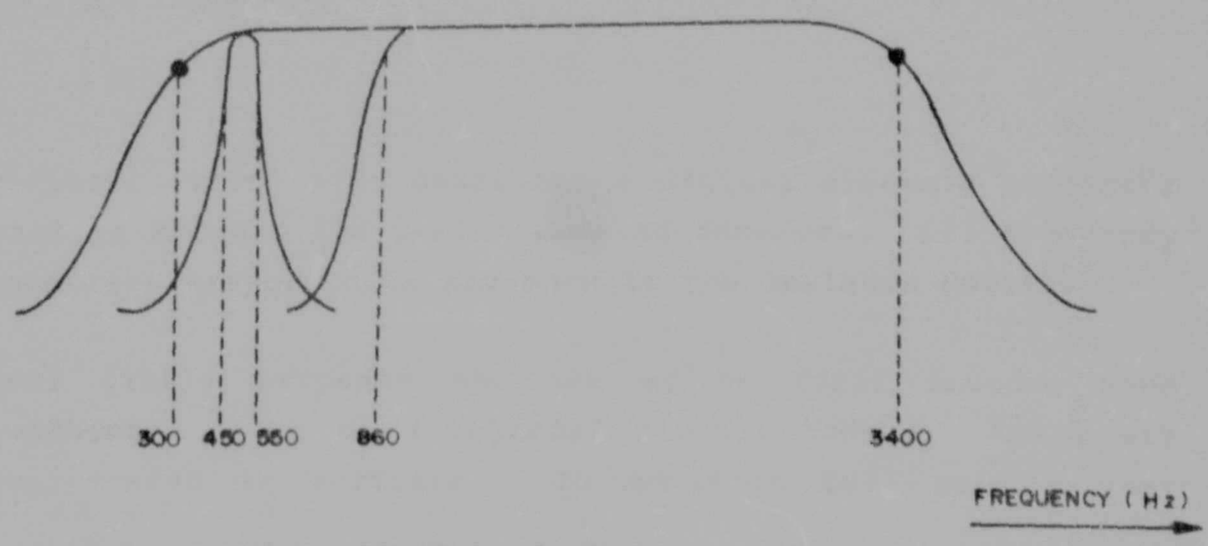

FIGURE 19 CHANNEL FREQUENCY SCHEMATIC

The above method requires signal generation and measurements to be done in the analogue domain. Hence the test is done on an encoder and decoder coupled in cascade. For separated performance measurements, signal generation and measurements are required both in the analogue and digital domains.

Molinari [1977] suggests a method for testing encoders by making use of digital filters. This is shown in figure 20. The stimulus used to excite the encoder is generated in the analogue domain.

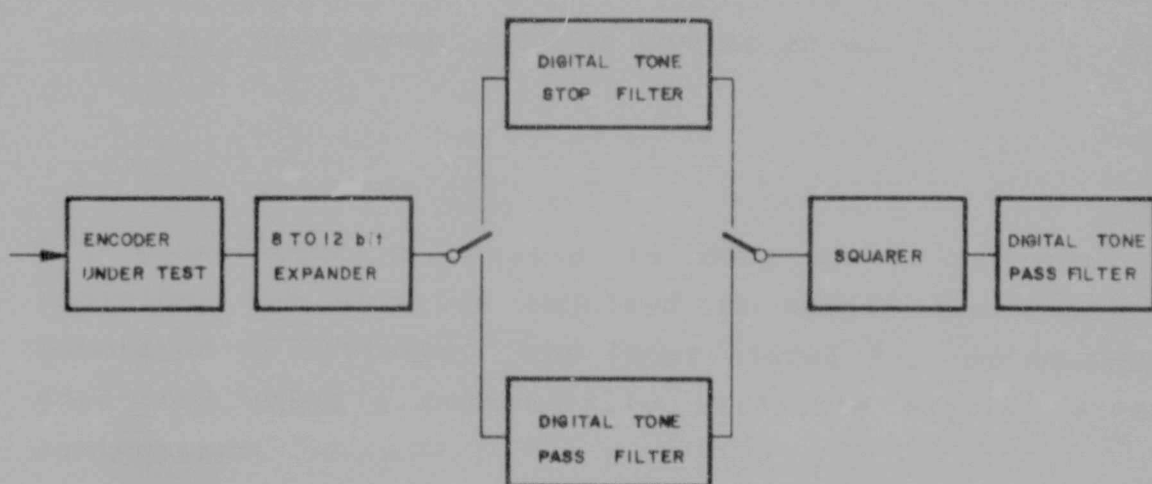

FIGURE 20 DIGITAL SIGNAL TO QUANTIZATION DISTORTION MEASUREMENT FOR ENCODER

Molinari [1976] also describes a digital sinewave generator used to measure the performance of decoders. All frequency selective measurements are done in the analogue domain.

Dell [1981] proposes the use of an ideal encoder when conducting tests on integrated circuit codecs. These are implemented in software. In addition Dell uses a Fast