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VARIABLE TIME-SCALE INFORMATION PROCESSING

A thesis submitted for the degree of

DOCTOR OF PHILOSOPHY

by

Mehmet Emin Binal

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Department of Electrical Engineering
University of Aston in Birmingham.

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SUMMARY

With the extensive use of pulse modulation methods in telecommunications, much work has been done in the search for a better utilisation of the transmission channel. The present research is an extension of these investigations.

A new modulation method, 'Variable Time-Scale Information Processing', (VTSIP), is proposed. The basic principles of this system have been established, and the main advantages and disadvantages investigated.

With the proposed system, comparison circuits detect the instants at which the input signal voltage crosses predetermined amplitude levels. The time intervals between these occurrences are measured digitally and the results are temporarily stored, before being transmitted. After reception, an inverse process enables the original signal to be reconstituted. The advantage of this system is that the irregularities in the rate of information contained in the input signal are smoothed out before transmission, allowing the use of a smaller transmission bandwidth.

A disadvantage of the system is the time delay necessarily introduced by the storage process. Another disadvantage is a type of distortion caused by the finite store capacity. A simulation of the system has been made using a standard speech signal, to make some assessment of this distortion.

It is concluded that the new system should be an improvement on existing pulse transmission systems, allowing the use of a smaller transmission bandwidth, but introducing a time delay.

CONTENTS

ACKNOWLEDGEMENTS

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I would like to dedicate this thesis to my dear wife whose continuous encouragement made this work possible.

CHAPTER 1

INTRODUCTION

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1. INTRODUCTION

Since the beginning of information transmission through band-limited channels, noise has been the major problem of engineers. The quality of communication systems is judged by their noise characteristics.

Any electrical signal corresponding to a message of some kind can be produced by using a transducer. An important problem of communication is finding a way to transmit this signal to a receiver so that it is as faithful as possible to its original form. This can be done simply by applying the signal in its analogue form to a twisted pair cable. But with this type of transmission as distance increases the noise level may become so high that the received signal is no longer usable.

In recent years, to overcome the noise problem a digital form of signal has been used. An example of this is the Pulse Code Modulation (PCM) system. In many ways this type of signal is similar to that used in telegraphy, which was the first type of electrical signalling ever to be used.

To form digital information, it is necessary to produce a digital message corresponding to its analogue counterpart. For this purpose a sampling, quantizing, and coding process is used. These processes have caused some impairments to the signal somewhat similar to amplitude and frequency distortions of analogue signals.

Although a digital signal is not able to give continuous information about the source signal's amplitude, a satisfactory quality can be achieved by using a suitable sampling law.

The digital form of signal is far more resistant to noise than any other analogue signal because of its two-level character. But the price paid for this is the necessity for a wide band-width of the transmission channel.

In recent years, great efforts have been made to use the available communication channels to their maximum possible signal capacities. This has been done in various ways with various systems.[1,5,9,13,29]

All transmission channels have been designed for transmitting the maximum possible frequency harmonics of the source signal. But most of the time, this provision is not necessary because of the varying characteristics of the input signal. Therefore, as far as the frequency bandwidth of the transmission channel is concerned, there is still plenty of spectral space for handling more information.

Present techniques are being used to utilize the capacity of the channel to better advantage. This, of course, is done by sacrificing the quality of transmitted signals. Every transmission system has a criterion for the acceptable quality of the received signal. Within the limits of this criterion it is possible to find the maximum amount of information which can be conveyed through a given channel. An example of achieving greater utilization of the channel capacity to better advantage is the system called multiplexing[5] .

With this investigation we looked for a system which could enable us to transmit a source information through a channel which would normally not be able to convey the same information without degrading its quality when using normal PCM techniques.

With PCM, the source signal is sampled at regular time intervals. Because of this, whether there is a change in source-signal amplitude or not, measurements are being made and sent in digitized form. If the amplitude has changed, a second sample would be carrying some useful information as well as the first one. But, if not, information in the second sample would be redundant. If redundancies can be replaced with some other information produced by another source, it is possible to multiplex these two and other systems [23].

With the system investigated in this research, sampling, quantizing and coding processes have been looked at from a different point of view. As useful information is produced only when a predetermined amplitude quantum change occurs in the source signal, then only at these times is a measurement made. Therefore, time gaps between two consequent quantized amplitude level changes are measured. So, the coded signal would contain information on measurements of the time rather than amplitude. This can also be quantized and coded as would amplitude measurements in the PCM system.

It is obvious that the occurrence of this data will be irregular. If it were to be transmitted as it occurred, the system would have little advantage over PCM. However, we can use a data regulator to take advantage of the system. For this purpose a dam-type store is planned on first-in, first-out basis. By doing so a constant rate of transmission of data from the transmitter is achieved.

When the input data rate of the store is higher than the transmission rate, the store will be filled up. As it is not feasible to occupy an infinite capacity store, from time to time there will be no place for the new data; the store will 'overflow'. Until the first vacant storage location occurs within the store all new data will be lost from the beginning of the overflow period. This characteristic of the system will involve a different type of impairment to the transmitted information from that encountered in other transmission systems.

When the input data flow rate is lower than the output data flow rate yet another problem occurs. This time the store will 'underflow'. When the store underflows the transmitted data does not contain useful information, in other words it will be redundant. A great deal of investigation has been done in this research to minimise the amount of redundant information and lost data.

In the experiments conducted the effects of the store capacity and transmitting frequency on the overflow and underflow has been investigated. Also, a statistical analysis of the system has been done using a standard speech waveform as a source-signal.

In this research the essentials of the idea have been investigated, and the relations between the system parameters investigated. It has not, in the time available, been possible to build a complete telecommunications system based on these principles. This awaits a further development. However, a system simulation has been prepared, enabling a subjective assessment of the distortion caused by overflow to be made.

CHAPTER 2

HISTORY OF THE PROJECT

CHAPTER 2

2.1 HISTORY OF THE PROJECT

Research on this Variable Time-Scale Information Processing (VTSIP) system was started in October 1971, when an examination of possible systems was made. It became obvious that whatever system was chosen, the dominant problem was that of information storage. It was therefore decided to devote some effort to the evaluation of the size of the store, as this represents the major part of the cost of the system. A search of the literature failed to reveal any information which would enable this to be done.

A simulation of the system was therefore attempted, using the University digital computer. Owing to the speed limitations of the computer circuits, it was impossible to carry this out in real time: a tape recording of a typical signal was therefore made, and played back at a reduced speed. Unfortunately, the resulting impairment in signal/noise ratio caused inaccuracies which cast grave doubts on the validity of the results of this exercise. After some months spent trying to improve this situation, it became clear that an alternative method of evaluating the store size requirements would have to be found.

The following few months were therefore occupied in devising and constructing apparatus for the analysis of speech signals described in Chapter 8. The operation of this equipment was satisfactory, and the data obtained from these experiments were used in the subsequent design of the system.

2.2 EXISTING PULSE MODULATION SYSTEM (PCM)

The system to be described has been developed as an extension of voltage will not be exactly the same.

Basically, sampling is the process of measuring the amplitude of a signal voltage in a very short period of time during which signal amplitude changes by a negligible amount. By sampling a signal voltage periodically it is possible to obtain information about the signal itself. A periodic sampling process is shown in Fig.2.1. According to the sampling theory 'a signal voltage containing no frequencies higher than $f_s/2$ can be completely determined by giving its amplitude at $1/f_s$ seconds aparts over the time axis'.[22,17] If we symbolize the sampled signal waveform as $f(t)\ rep_{T_s}\ \delta(t)$ where f(t) is the signal voltage and rep_{T_s} (δt) is the repetetive $\delta(t)$ with T_s periods, after applying this signal to a low pass filter we obtain:

$$G(t) = \left[f(t)\ rep_{T_s}\ \delta(t)\right] * \text{sinc}(\omega_s t) \qquad 2.1$$

$$\omega_s = 2\pi f_s$$

where $\text{sinc}(\omega_s t)$ is the ideal low pass filter characteristic. The spectrum of this expression can be written :

$$G(\omega) = rep_{\omega_s}\ F(\omega)\ .\ \text{rect}\ (\omega/\omega_s) \qquad 2.2$$

where $F(\omega)$ is the Fourier transform of f(t) and rect (ω/ω_s) is the rectangular function. From this equation it is evident that $G(\omega)$ vanishes for higher frequencies than ω_s. Therefore, $G(\omega) = F(\omega)$ and $G(t) = f(t)$ can be written. This clearly shows that a signal voltage is determined by its samples. In other words by transmitting $2f_h$ independent samples per second as long as the signal exists we can reproduce the signal almost without distortion. This is, of course, assuming the signal is ideally band limited at f_h.

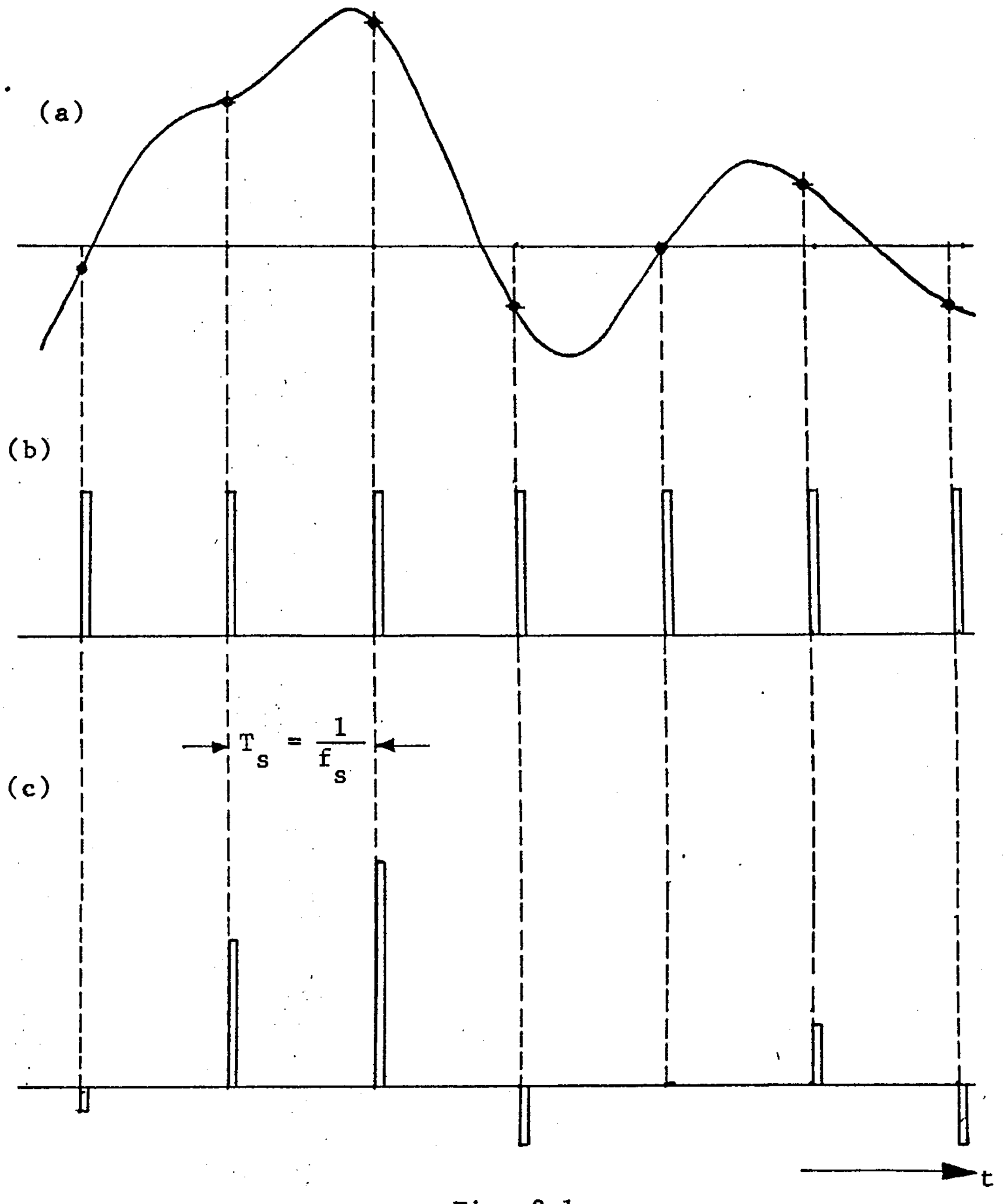


Fig. 2.1

PERIODIC SAMPLING PROCESS

(a) Signal
(b) Sampling pulses
(c) Amplitude modulated sampling pulses.

Each sample taken from the signal may be any value within certain limits. Sending these amplitude modulated pulses will not be any advantage if transmitted in this form, as they will still have the disadvantages of the analogue signals. The first thing to reduce the amount of information needed to a minimum degree, a quantizing process is carried out. By doing so, each sample amplitude is approximated to the nearest quantized amplitude level. This approximation introduces an error which is usually called 'quantizing noise'[4,5]. Although this error is correlated to the signal, if there is a very small difference between amplitude quantum levels the correlation becomes insignificant. A quantizer is simply a circuit having a stepped transfer characteristic, as shown in Fig.2.2. The amplitude of the steps are determined by the subjective quality tests. In Fig.2.3 a uniform quantizing process is illustrated. For some types of signal a uniform characteristic may not give better results, and some other types of characteristics can be employed, e.g. logarithmic quantizing[5,15,28].

Before quantizing the signal, maximum deviation of its amplitude from the origin must be known. According to this knowledge, if uniform quantizing is employed, this level is divided into m equal quantum levels. Therefore, a set of continuous levels are represented by only some of m quantum levels. If m is large enough quantizing noise becomes very small. For non-periodic signals the spectrum of the quantizing noise is very wide and widens even more by the increasing number of levels. By the introduction of non-uniform quantizing characteristics it is possible to obtain better quantizing noise performances .

The squared quantizing error can be found from :

$$e^2 = \frac{1}{12} a^2$$

where a = amplitude of a unit quantum.

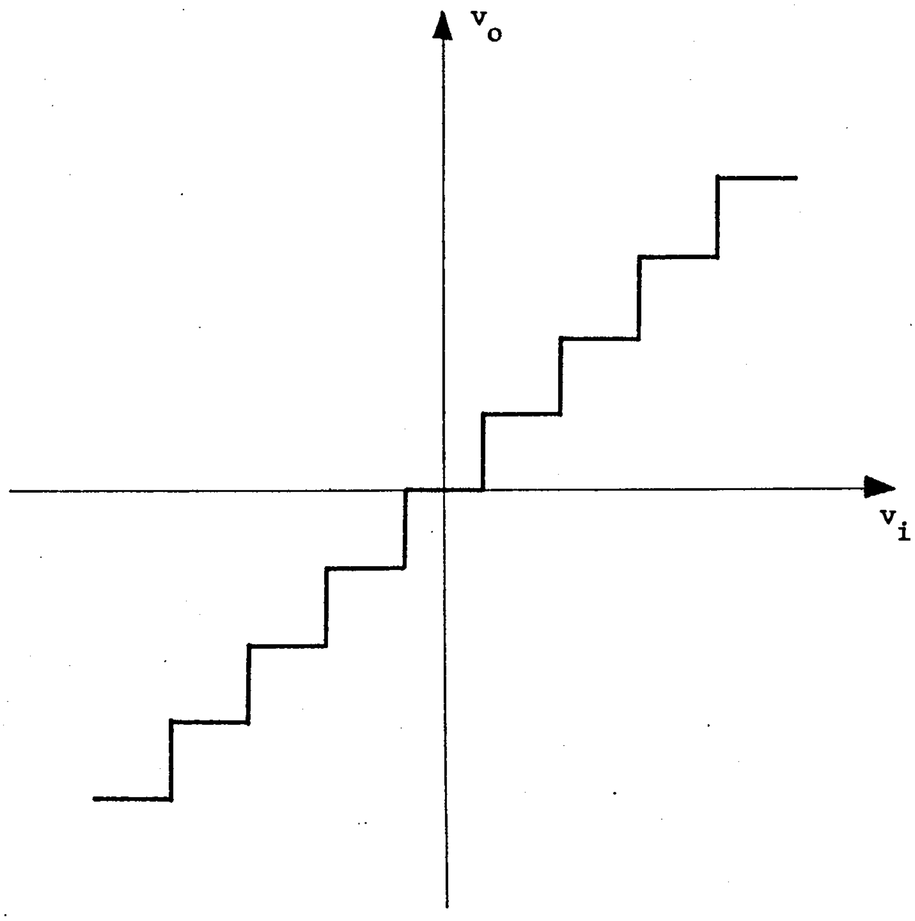


Fig. 2.2. QUANTIZER CHARACTERISTIC.

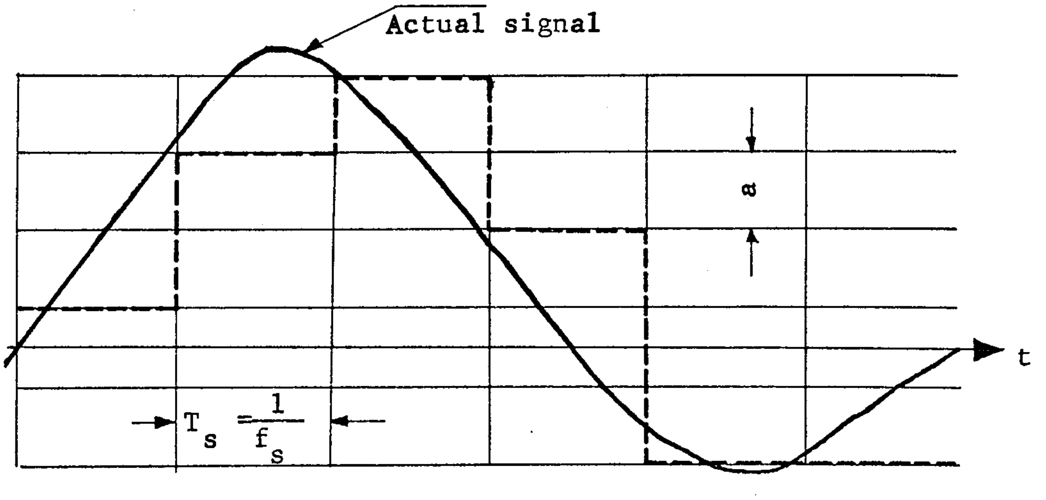


Fig. 2.3. QUANTIZED-SAMPLED SIGNAL. (Shown in dotted lines)

As the number of quanta is high error signal will have negligible correlation to the original signal. The spectrum of the error signal, therefore, will be very wide approaching any likely sampling frequency. By the sampling process this spectrum will be folded back and the noise contents corresponding to the signal band will be considerably higher than expected. (See Fig. 5.1). If the sampling frequency is very high, in the limit this will correspond to the unsampled quantizing noise situation. But if we increase f_s we will need a very wide channel bandwidth.

After the quantizing process, each sample is represented by one of m quanta levels. By using binary counting methods the levels having a value between 0 and m can be shown in the form of pulse groups. Each group being coded to a number corresponding to one of these quanta levels. By using binary numbers it is possible to code m levels as :

$$m = 2^n \tag{2.3}$$

Therefore, information content of a code representing a sample of maximum deviation becomes :

$$I = \log_2 m = n \text{ bits} \tag{2.4}$$

To calculate the occupancy of the transmission channel in worst case situation, information per second can be written :

$$\text{Channel occupancy} = f_s \cdot n \quad \text{bits/sec.} \tag{2.5}$$

where f_s is the sampling frequency. As we choose $f_s \geqslant 2f_n$, channel occupancy becomes :

$$\text{Channel occupany} = 2f_h.n \quad \text{bits/sec.}$$

Transmission medium must be capable of conveying this information. Obviously, a signal rarely approaches the amplitude levels as high as this. But as the possibility is still there, system designs must be done considering the worst case.

It will be appreciated that in the PCM system the transmitted pulse group frequency is constant. There will, therefore, necessarily be a waste of potential channel capacity during those times that the signal amplitude is not more than a few quanta levels. Many attempts have been made to take advantage of this potential channel capacity by using numerous methods, rather than wasting it.[4] This investigation represents a further attempt for the purpose of producing information which can be conveyed through a channel having less potential capacity.

CHAPTER 3

DESCRIPTION OF THE PROPOSED SYSTEM

CHAPTER 3

3.1 DESCRIPTION OF THE PROPOSED SYSTEM

Unlike other modulation systems, the proposed system does not produce amplitude information about the source-signal which is to be transmitted. Therefore, it is essentially different from the rest of the modulation systems.

The processed information, corresponding to the source-signal is a measure of time-length which is spent between two consequent quantized amplitude levels. Every time the signal amplitude increases or decreases by a unit quantum this information is being produced. Therefore, information is produced irregularly as the source-signal is random. But it is worth mentioning that each bit of information determines a definite change in the source-signal amplitude. If we make a comparison with the other digital communication systems, it is obvious that the sampling technique used here is not like the others. With other systems samples are taken periodically to produce information about the amplitude change of the input signal within this period. This sampling period is, of course, suitable enough to get an acceptable quality at the receiving end of the communication link.

Information processed as explained in the above paragraph is contained in a digitally coded word. Each word, therefore, is produced after each amplitude quanta change of the source-signal. To regulate this irregular information flow the use of a store is necessary. If a suitable size store is employed, then by pulling out stored information with a constant rate, this regulation process can be realized. The pulling out rate, must be equalled by the average rate of information processed. With the proposed method it will be possible to convey this information through a communication channel without occupying it unnecessarily.

The system, therefore, consists of a circuit which derives information, 'information processor', and a store. The information processor is basically a level sensing detector which produces a pulse every time a unit quanta amplitude change occurs in the source-signal, and a time interval measuring device to measure the time lengths between the consecutive pulses. The latter device also quantizes and codes this time-length measurement. The store is a 'silo' type store, as commonly called. Registered information is taken out by the transmitter at a constant rate. At the receiving end similarly, there is a store and the circuits which convert the coded time-length information to voltage. The design of the circuits and store is very dependent upon the source signal. For high values of signal the data rate will be very high. Therefore, the circuits must be able to operate at these speeds.

At the receiving end there is another store which registers the regularly received information from the transmitter. Each pulse group which contains the time-length information, directs to the transmitter to hold the present amplitude level, and at the end of this period the next pulse group gives another directive about the time-length of the next amplitude level. (See Fig.3.1).

For a store, it is possible to use various types of devices such as paper tapes, magnetic cores, magnetic tapes, or semi-conductor elements. The choice of store will depend upon the type of source signal, and is likely to be greatly influenced by

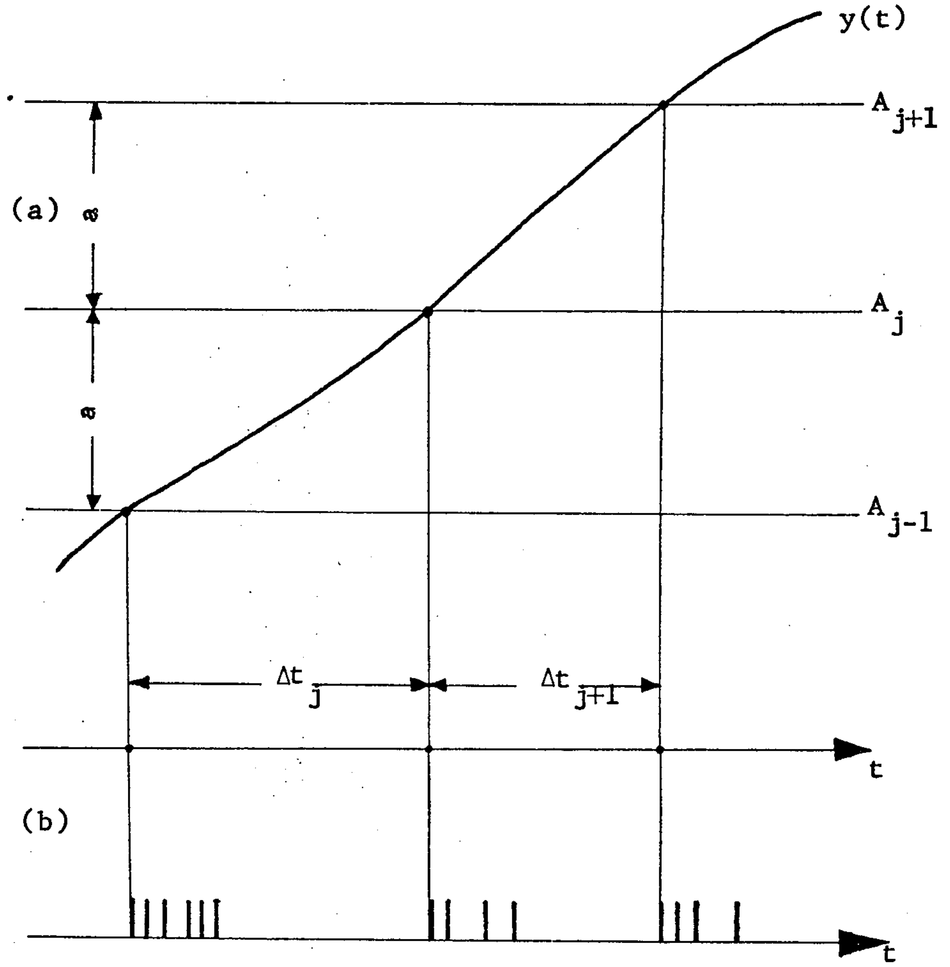


Fig.3.1 (a) Amplitude changes of the signal
(b) Information carrying pulse groups.

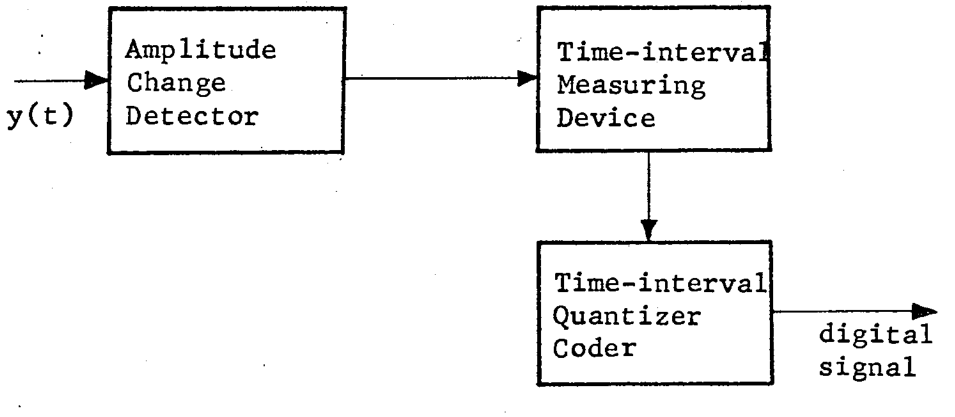


Fig.3.2 INFORMATION PROCESSOR

economic considerations.

3.2 GENERAL REQUIREMENTS OF THE SYSTEM

Once the maximum slope of the input signal is known, a device which is capable of producing time-length measurements of unit amplitude quantum changes of the source-signal can be designed. By carefully studying the characteristics of the signal to be transmitted, all necessary design information can be obtained.

It is convenient to call this device the information processor. (Fig. 3.2). It consists of an amplitude change detector, a time-length measuring device, and a time-length quantizer and coder.

The amplitude change detector senses the predetermined amplitude level changes of the source signal. It produces a pulse every time the input signal crosses one of these amplitude levels. The device has two outputs indicating the changes in either direction. (Fig. 3.3). These outputs are fed into the time measuring device, which measures the time-interval between two consequent pulses which are produced by the previous circuit.

The result of this analogue measurement is then quantized and coded in digital form, ready to be transmitted. Each digitally coded word contains information indicating a unit amplitude quantum change to last the length of time which is encoded within itself.

Depending upon the source signal there may be long silence

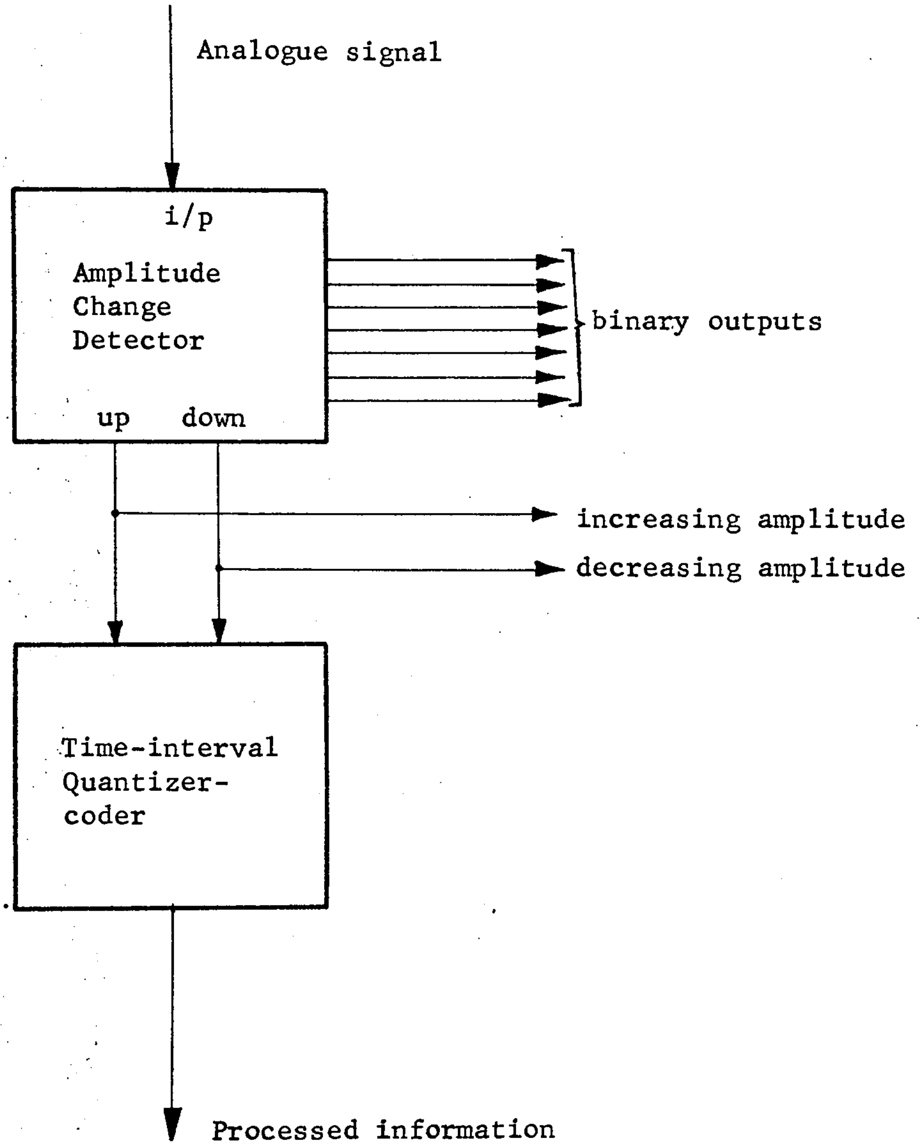


Fig.3.3. DIGITAL OUTPUTS OF THE AMPLITUDE CHANGE DETECTOR.

gaps as well as very frequent amplitude changes. As one of our most important goals, information must be transmitted with a constant rate. Therefore, as mentioned before, a store serving as an information regulator is used. At the input of this device, digitally coded words containing information occur randomly. The store, which is designed on first-in, first-out basis conveys this information to the transmitter at a constant rate. If there is a higher rate of information flow at the input of the store, than at the output, its occupancy increases. As there will be a limit to the capacity of the store to be used, continuity of such a situation inevitably causes an overflow condition.

On the contrary, if there is a higher rate of information-flow at the output of the store than at the input, there will soon be no information in it to be conveyed to the transmitter. This creates the underflow situation. After that instant, the rest of the signals will be of no use as they will not contain any information. It is clear that the higher the store capacity, the lesser the possibility of underflow as well as overflow.

The use of the store must cause a delay in the system, which delay will increase as the store capacity increases. For such a delay there are practically permissible limits for any particular communication system. This represents one limit to the size of the store.

By the addition of a store and a transmitter to Fig.3.1, the complete system block diagram can be drawn. Regularly received information is fed to the store as at the transmitting end. But information is pulled-out irregularly from the store under the instructions given on the decoded time-length information. A detailed explanation of the transmitter and receiver is given in Sections 3.2.1 and 3.2.2.

3.2.1 Transmitter

As explained in Section 3.1 there are two main parts of the transmitter, which are the data processor and the store.

A block diagram of the transmitter is shown in Fig.3.4. As explained previously the amplitude-change detector produces a pulse every time an amplitude level is crossed in either direction. Then, a flip-flop which is triggered by these pulses opens the gate of the counter. This counter starts to count until stopped by the next pulse reversing the flip-flops state. Immediately a second counter starts to count the unit pulses until a further pulse arrives. At any time each of these counters contains, in quantized form, time-length information about two consequent amplitude level crossings.

While one counter is counting the unit pulses, information reserved in the other counter is registered to the store. A multiplexer conveying information in sequence is used before the store.

The unit pulse generator produces the pulses with a constant repetition period corresponding to the unit of the time-scale. Therefore, this period represents unit time-quanta. By counting these pulses, information will be quantized automatically. As a binary counter will be used, its output will represent a binary coded word containing the quantized time-length information we need. This method of quantizing and coding is quite basic and easy to implement.

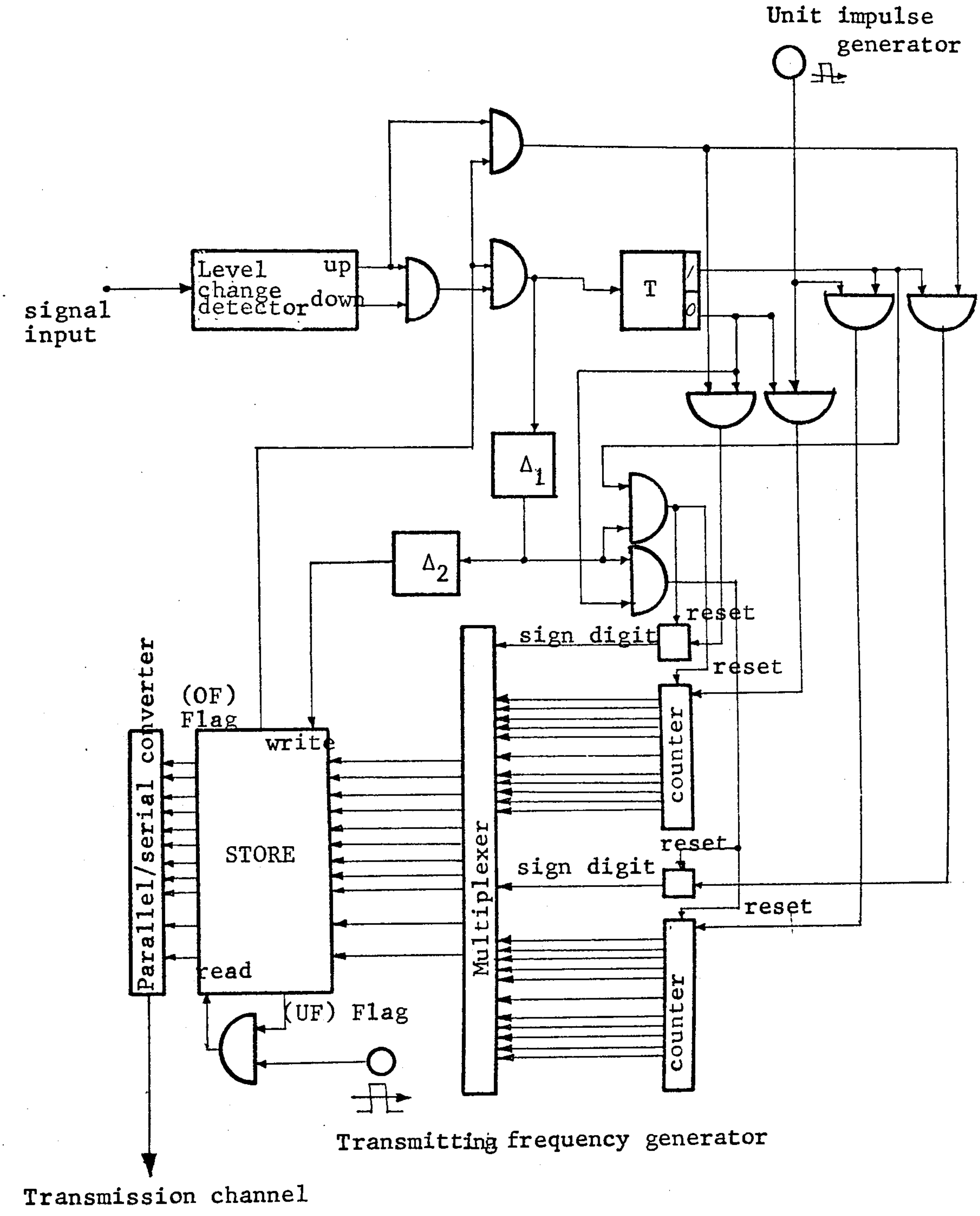


Fig.3.4 TRANSMITTER BLOCK DIAGRAM

As the data flow increases the store will become full after a time, and this condition will be indicated by an overflow flag. Then the overflow flag will be low, and a gate will block the input of the flip-flop preventing a further change of state. This situation will continue until there is an available space in the store to register new information, so the flag output will be high again to open the gate, permitting the flip-flop to change its state.

When the data flow decreases at the input of the store, this time an underflow condition will occur. This will be indicated by an underflow flag at the output of the store. To avoid sending useless information it is possible to block the output of the transmitter during the times this conditions persists.

To indicate the polarity of the movement of amplitude, an extra digit is included in the digital code-group transmitted. This is sent as a '1' when the increment is positive.

3.2.2 <u>Receiver</u>

A block diagram of the receiver is shown in Fig.3.5. It is first necessary to convert the received serial code groups to a parallel form. When the word length is complete a write pulse transfers this information to the store. In the meantime a counter, similar to the one used in the transmitter, counts up the unit pulses generated by the unit impulse generator. When they are equal, the comparator produces a pulse which, on being integrated by a simple integrator circuit, produces a step signal at the output lasting until the next pulse comes. The same comparator output is applied to the reset input of the counter, as well as to the 'read' input of the store, after a delay sufficient to permit

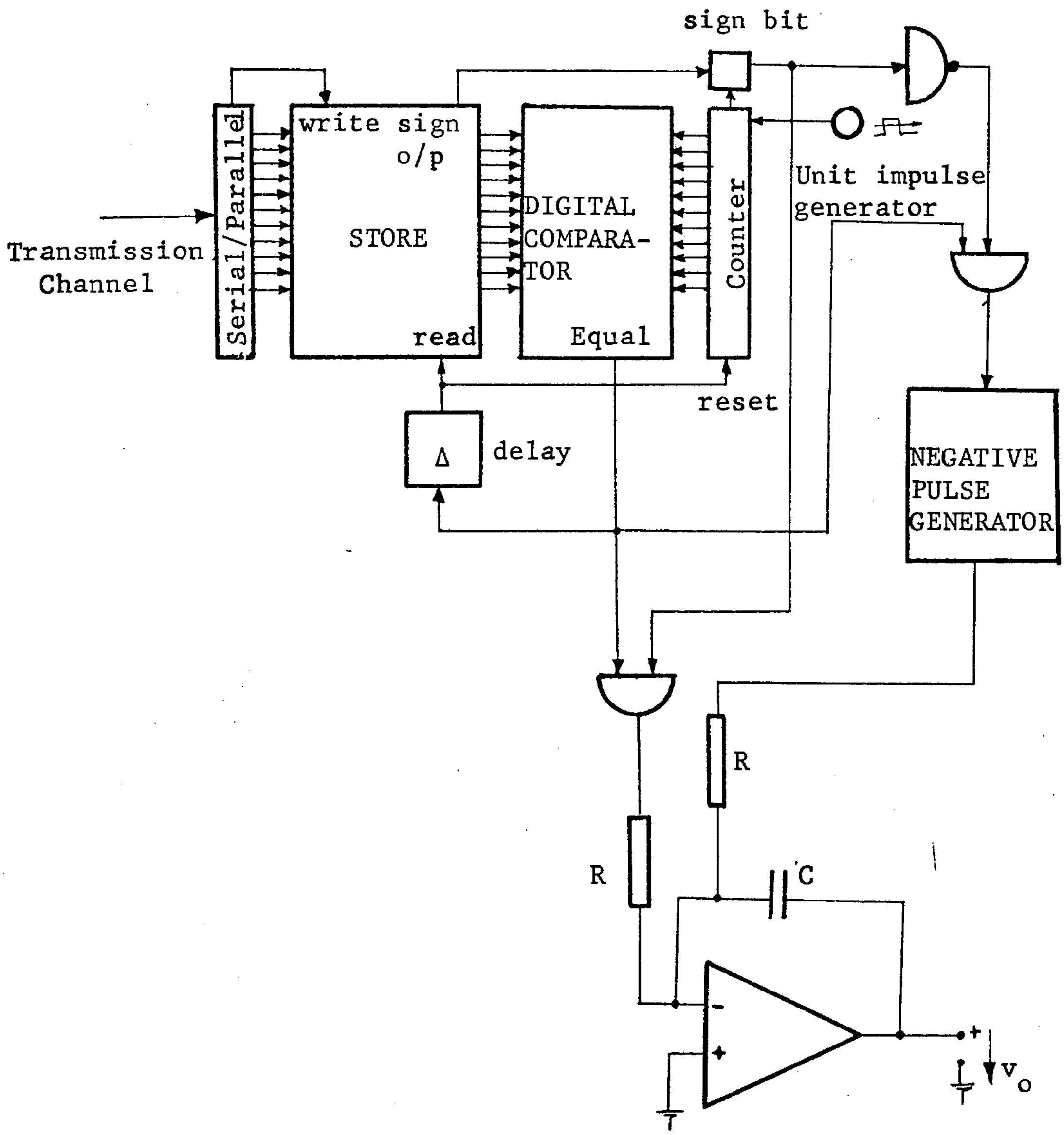


Fig.3.5 RECEIVER BLOCK DIAGRAM

charging of the integrator capacitor. This action applies for the positive slopes of the source signal. This delay is necessary because without the delay, counter will be reset without being compared with the information, which is at the output of the store.

For the negative slopes the sign indicator digit contained in the received word becomes '0' or 'low', blocking the positive pulse path. Instead, when the digital output of the counter equals the received word, comparator output is triggered. This output pulse from the comparator excites a negative pulse producing circuit. The negative pulses, when applied to the integrator, decreases its amplitude by a unit step. Therefore, the integrator output gives a stepped output signal proportional to the source signal. (See Fig. 3.6).

There will be distortions of the waveform because of the amplitude quantization. This is investigated further in Chapter 5. Store overflow is another source for waveform distortion, which is investigated in chapter 4.

3.3.1 Time Coding Methods

As explained in Section 3.2.1, the transmitted signal consists of pulse groups which represent a measure of time intervals. The shortest time interval will occur under conditions when the signal voltage has the maximum rate of change. Assuming this signal to be quantized in equal increments of voltage, the greatest rate of change will occur with a signal consisting of a sinusoidal voltage having the maximum amplitude and frequency for which the system is designed, and measured at the point where the voltage passes through zero value.

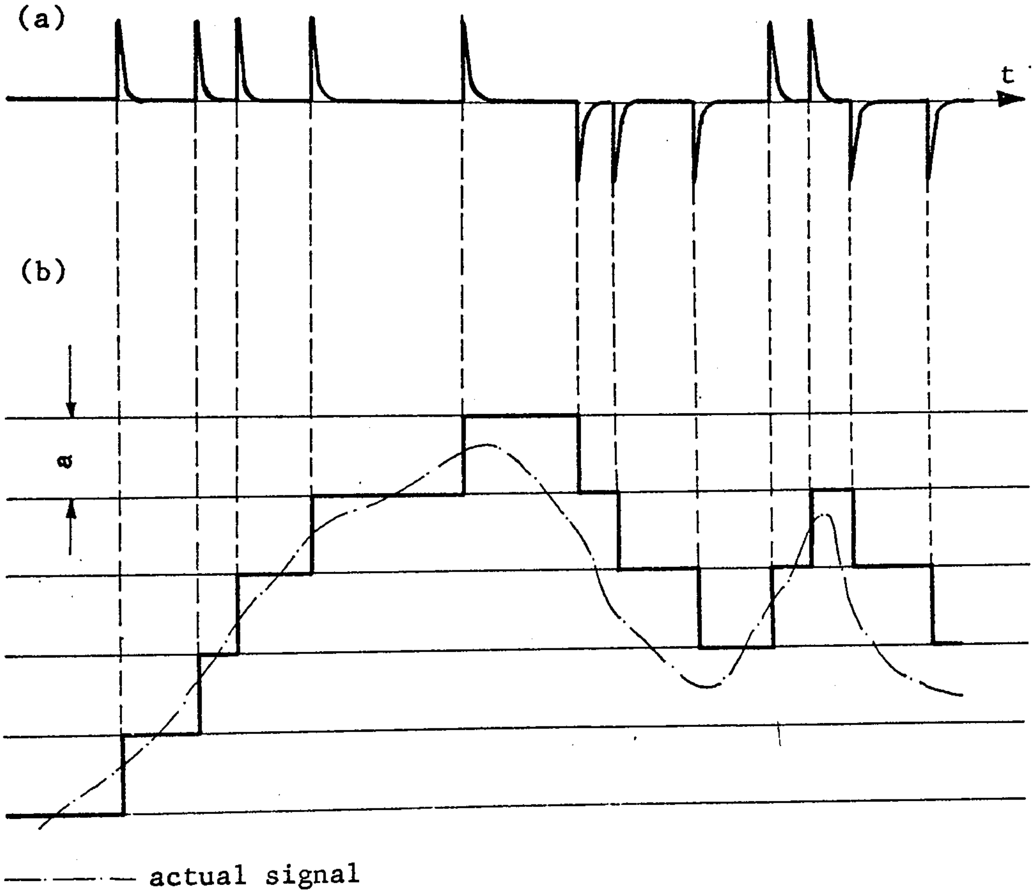


Fig. 3.6 RECONSTITUTION OF THE SIGNAL

(a) Pulses at the input of the integrator

(b) Reconstituted signal.

The least rate of change is less easy to evaluate. During periods when the signal voltage is not changing, (for example, when it is zero), a measurement of time interval would require a very large transmitted pulse group; the accuracy of measurement in such a case could be reduced without too great a sacrifice in the received signal quality. This is discussed in Section 3.3.2. However, when there is a lower limit to the frequencies contained in the signal, the smallest rate of change sensed by the system will occur with a voltage of this lowest frequency, having a peak-to-peak amplitude equal to one quantum.

3.3.2 Measurement of Time Intervals

Since the transmitted signals representing the time intervals between successive samples of the signal voltage are in the form of pulse groups, it would appear that a suitable method of time measurement might be to count the number of pulses generated by an oscillator during the interval, and to use the digital output of the counter to determine the configuration of the pulse group. It would be possible to adjust the increments of time measurement by variation of the oscillator frequency, should this be desirable.

A diagram of a time measurement system working in this manner is given in Fig. 3.7. Each time the signal voltage crosses a quantum level, the flip-flop A is triggered, allowing pulses from the oscillator B to reach the counter C. The output of this counter is therefore a measure of the time interval elapsing between successive voltage amplitude quanta. The fact that this measurement of time is necessarily quantized gives rise to a further source of error in the reconstituted signal, which will henceforth be called the 'time quantum error'.

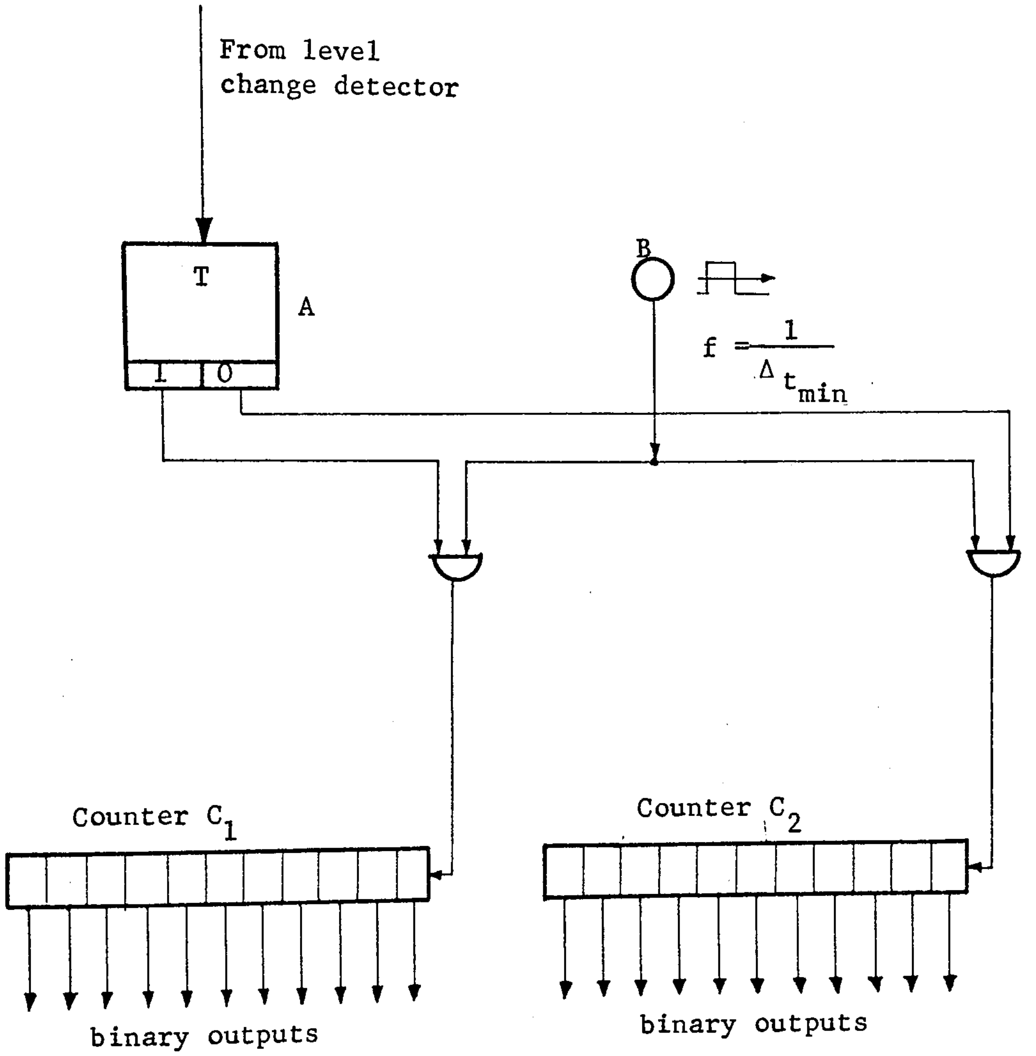


Fig. 3.7. TIME INTERVAL MEASUREMENT SYSTEM.

To be able to measure the shortest time interval with a certain error, the oscillator must have a suitably high repetition frequency. The shortest time interval will be measured when the rate of change is maximum. As the oscillator frequency increases, the measurement will be finer. This naturally causes an increase in the rate of information produced.

Long intervals of time also have to be measured and transmitted using the same type of pulse groups. The longer the time interval the greater the length of words containing information. This also results in an increase of information rate.

The direct counting method described above has an obvious disadvantage. If the oscillator frequency is arranged to be sufficiently high that the time quantization error is acceptable with signals having their greatest rate of change, then for signals with a small rate of change the number of pulses to be counted between successive quanta will be very large. This will require an excessive number of counting elements, and would represent an unnecessarily high accuracy of time measurement. There are many possibilities for the modification of the system, some of which are discussed later in this thesis. However, if the measured time interval is long it is unlikely that a much shorter time interval will follow it, considering that the signal is a stationary random signal.

Before deciding about the code-length to be used for coding the pulse groups which carry information about the time intervals, there is more to be said about the characteristics of the signal for which the system is designed. This investigation also helps to establish the accuracy limits of our measurements.

3.3.3 The Coding Process

In general, to decide about the source-coding process of the system the first thing to be considered is the bandwidth of the source signal. As explained in section 3.3.1 the high and low end of the frequency band gives us an idea about the extreme rates of change. Assuming that we have quantized the amplitude in m equal increments, the maximum rate of change occurs with a signal of maximum amplitude and frequency. If the lowest frequency is f_1 and the highest frequency is f_h of the signal, bandwidth w can be written:

$$w = f_h - f_1 \qquad 3.1$$

let a sinusoidal y of maximum amplitude and frequency be :

$$y = \frac{A}{2} \sin 2\pi\, f_h t \qquad 3.2$$

where $A = m \times a$

$a =$ unit step voltage.

The slope of this voltage waveform will have its maximum value around $t = 0$. Therefore the shortest time interval Δt_{min} which can be written as follows:

$$\Delta t_{min} = \frac{1}{2\pi f_h} \text{ arc sin } \frac{2}{m} \qquad 3.3$$

The maximum time interval is measured when the lowest frequency signal crosses only the first level and can be expressed as :

$$\Delta t_{max} = \frac{1}{2 f_1} \qquad 3.4$$

All other time interval measurements will vary within these two limits. Obviously, there will be time-intervals which are much longer than Δt_{max}, e.g. when there is no signal coming from the source. The counter in this case may overflow. Therefore the resulting measurement will be wrong. But because of the characteristics of the system this mistake in measurement will not be so effective.

If Δt_{min} is accepted as the unit time quantum a code length of $1'_c$ can be calculated as :

$$1_c = \log_2(\Delta t_{max} / \Delta t_{min}) = \log_2 \frac{\pi f_h}{f_1 \times \arcsin \frac{2}{m}} \qquad 3.5$$

Assuming a long silence-gap has occurred and the counter has re-cycled k times, exactly $(k \, . \, 1_c)$ bits will be missed. There are two possibilities:

(1) If the store in the transmitter was not emptied, in this case if Δt_i is the actual time interval, $(\log_2(\frac{\Delta t_i}{\Delta t_{min}}) - k \, . \, 1_c)$ bits will be transmitted. Therefore, the time interval measurement error will be :

$$\Delta t_e = k \cdot \Delta t_{max} \text{ secs.}$$

(2) If the store in the transmitter was emptied, and the state of the store occupancy was S_i when the silence gap has started, assuming the transmission rate is R_o, error will be:

$$\Delta t_e = \Delta t_i - \frac{S_i}{R_o} \text{ secs.}$$

This time interval measurement error may become a disturbing factor. This has to be individually investigated for the systems concerned.

For our system to be competitive $(1_c \, . \, R_o)$ must be less than any other digital systems transmission rate. In this research it is accepted that Δt_{min} is the unit for the time quantization.

To indicate the direction of the amplitude change, an extra bit is added to the coding system, making the total code word length $(1_c + 1)$ bits.

CHAPTER 4

SYSTEM PARAMETERS

CHAPTER 4

4.1 SYSTEM PARAMETERS

Variable time-scale information processing consists of two different steps. The first step is to convert the source signal voltage into a series of irregularly occurring pulse groups containing information about the time intervals. The second step is to regulate the flow of this information and transmit it to the transmission medium.

Within the first step, a sampling quantizing and coding process is implemented. The sampling is not in the sense of the sampling process used in the PCM system, but is an 'amplitude sampling'. This means that instead of making a measurement of the instantaneous voltage with regular time intervals, we are measuring the time spent between two adjacent quantized amplitude levels of the voltage. So the amplitude quantization is again used in the same manner as the PCM system. In addition to the amplitude quantization, measured time-intervals are also quantized in order to be able to digitize this value with a pulse group. Each pulse group is coded in a form which is most suitable for the system to be used. Therefore a pulse group containing the time-interval measurement is produced with such a process.

This main difference of our system from the other telecommunication systems, led us to search for some other properties of the source signal.

It is obvious that the flow of information will increase with an increase in the number of amplitude levels. This will give a higher signal to quantization noise ratio, although it necessitates the use of a larger store. Therefore the store occupancy is a direct function of the number of levels used. Economic and practical considerations always put a limit to the size of the store needed. Subjective quality tests and calculation of the signal to quantization noise ratio give us an idea about the minimum number of amplitude levels which must be used. It was not intended to make subjective quality tests of this kind, as some other parameters were considered more important; the available time has been spent in the investigation of these, and a mathematical approach to finding the quantization noise level of the system. This level is not only dependent upon the number of amplitude levels, but on the time quantization as well. Therefore, the error caused by the time quantization must be taken into account.

The store occupancy is also a very important system parameter. Information rate increase causes the store occupancy to increase. When the store has no more places to register any more information, overflow will inevitably occur, until such time that the transmitter sends one more pulse group from the store. The pulse groups which are produced by the information processor will be lost as long as the overflow occurs. This loss of information will cause an 'overflow distortion'. As this does not occur very often and its period varies randomly it is not possible to formulate it in a meaningful style. But statistical investigation of the source signal may give us some idea. So far, transmission frequency is not considered to be an important parameter. Since the store size is limited, then with different transmission rates (R_o) different overflow distortions will occur. Namely, the store occupany is a function of R_o as well as number of levels (m).

In this chapter, an investigation for the system parameters is undertaken. The essentials of the theory, in general, are put forward. A practical verification of the theoretical results is made for a speech signal and is given in Chapter 8.

4.2 RELATION BETWEEN STORE CAPACITY AND TRANSMISSION FREQUENCY

The feasibility of the prepared system can be judged by the calculations of the store size to be used. Therefore, this aspect must be the most important of all.

If the statistical properties of the source signal are known, we can derive information about the rate of change and therefore the input rate of the store from this. We intend to calculate the relation of store occupancy function with time. But for signals like speech, information flow rate will be random. Therefore we cannot write a definite mathematical function in such a case. What we do is, (assuming the source signal is a stationary random process), to try to find the store occupancy density function by using the first derivative probability density function of the source signal. This mathematical approach has been done with the practical considerations at the back of our minds. Some simplifications and approximations have been made whenever appropriate, without doing any harm, hopefully, to the realities of the system.

Whenever the source signal voltage passes through one of the m predetermined amplitude levels, a pulse group indicating the time interval between two successive voltage levels is stored. A pulse group is called a 'word' which carries this information. In the meantime the transmitter periodically pulls out a word from the store. Therefore the rate of store occupancy change is the difference between the input and output rates. Store occupancy is not a continuous function, as the pulse groups are in discrete form. For the convenience of the calculations, store occupancy is assumed to be a continuous function, rather than a stepped function, which it actually is.

If Δt_i seconds has passed between two consecutive level crossings, and the state of the store occupancy is S_{i-1}, and the transmission rate is R_o, the store occupancy for the present state S_i is

$$S_i = S_{i-1} + 1 - R_o \cdot \Delta t_i \qquad 4.1$$

If Δt_i is a long time interval, so that the store is emptied by the transmitter which pulls out information with the rate of R_o, then the underflow condition should occur. For this situation we can write

$$S_{i-1} + 1 \leqslant R_o \times \Delta t_i \qquad 4.2$$

If $S_i > S_{i-1}$ for successive level crossings, and if the store size S_{max} is exceeded, an overflow condition will occur. This means, information produced from this moment on is lost, causing a distortion to the signal. For small sizes of store this condition may occur quite frequently, and the result may not be acceptable. Therefore S_{max} has to be chosen after a very careful consideration. For every source signal there will be another S_{max}.

As seen from the equation 4.1, the transmission rate directly affects the store occupancy. Low rates of transmission cause frequent overflow, but on the other hand, high rates of transmission cause frequent underflow. Therefore, there must be a transmission rate which causes acceptable overflow and underflow conditions for a particular store size. We will call this the optimum transmission rate.

It is evident that as the transmission rate increases this would cause proportionally higher underflow than overflow. On the contrary, if the transmission rate decreases this would mean proportionally more overflow than underflow.

If in the interest of received signal quality, we sacrifice potential channel capacity, by introducing underflow, we can increase the transmission rate. Underflow only means occupying the transmission channel unnecessarily. But this, being one of the most important drawbacks of PCM, is not desirable. As we cannot allow too much overflow we have to find an optimum store size and transmission rate for this particular system.

Since we can find out the statistical properties of the source signal, we can develop a theory to find the store occupancy density function. Using the result, it is possible to find statistically optimum store size and transmission rates.

In the next sections calculation of the general form of store occupancy density function is made.

4.3.1 Store Occupancy Density Function

In this section, a theoretical analysis has been made of the calculation of the store occupancy density function.

For simplicity the store is assumed as infinite, although it is not, of course, possible to use an infinite size store in practice. By observing the store occupancy of certain intervals and taking the time average of those observations, it is possible to produce the steady-state probability density function of the store occupancy.

In these calculations, store occupancy is assumed to be a continium of states rather than a number of discrete states, which it actually is. This assumption was made for the mathematical convenience.

The state of the store occupancy is dependent on the previous state only. Therefore this relation is first order Markovian[2,6,14]. Assuming that $S_{i-1} \geq 0$ the next store occupancy is dependent on the added data to the store in the period of Δt_i, which is :

$$\Delta S_{i+1} = 1 - R_o \times \Delta t_i \qquad 4.3$$

If we write the conditional probability distribution function for store occupancy:

$$P(S/S_i) = P_1(S, \text{Store did empty}/S_i)$$
$$+ P_2(S, \text{store did not empty}/S_i) \qquad 4.4$$

Obviously, if the store emptied, store probability distribution is equal to that of ΔS_{i+1}. But if the store did not empty, this time we must consider both S_i and ΔS_{i+1} together. In the above equation S_{i+1} and S_i are replaced by S and S_i respectively. If we can find $p(S/S_i)$ by using a well-known relation from probability theory,

$$p(S) = \int_o^\infty p(S/S_i) \cdot p(S_i) \cdot dS_i \qquad 4.5$$

If $R_o \times \Delta t_i = S_i{+}1$ occurs, S will be zero at the end of Δt_i period. That means that an underflow condition may occur, if the Δt_i period has lasted longer than $1/R_o$. Therefore, $\Delta t_i \geq \frac{S_i+1}{R_o}$ is the condition for the underflow.

It is known from equation 4.1, replacing S_{i+1} with S :

$$S = S_i + 1 - R_o \times \Delta t_i \qquad 4.6$$

So, where $\Delta t_i < \frac{S_i+1}{R_o}$ equation 4.6 can be written.

To find the conditional probability function of the store occupancy we must integrate $p(\Delta t_i)$, in the appropriate region of the Δt_i axis.

as shown in Fig.4.1 :

$$P_1(S_1 \text{ store emptied}/S_i) = \int_{\Delta t_i = \frac{S_i+1}{R_o}}^{\infty} p(\Delta t_i) \cdot d(\Delta t_i) \qquad 4.7$$

For the remaining parts of the axis as the store is not emptied:

$$P_2(S_1, \text{ store did not empty}/S_i) = \int_{\Delta t_{i\,min}}^{\Delta t_i = \frac{1+S_i}{R_o}} p(\Delta t_i) \cdot d(\Delta t_i) \qquad 4.8$$

By combining the equations 4.7 and 4.8 we can write:

$$P(S/S_i) = \int_{\Delta t_{i\,min}}^{\Delta t_i = \frac{S_i+1-S}{R_o}} p(\Delta t_i) \cdot d(\Delta t_i) \qquad 4.9$$

To find the conditional probability density function for that case :

$$p(S/S_i) = \frac{dP(S/S_i)}{dS_i} = \frac{1}{R_o}\, p_{\Delta t}\left(\frac{S_i+1-S}{R_o}\right) \qquad 4.10$$

Finally by substituting equation 4.10 into equation 4.5,

$$p(S) = +\int_0^{\infty} \frac{1}{R_o}\, p_{\Delta t}\left(\frac{S_i+1-S}{R_o}\right) \cdot p(S_i) \cdot dS_i \qquad 4.11$$

can be found, where $0 \leq S < \infty$.

Until now, to our knowledge, because of the different characteristics of the other telecommunication systems $p_{\Delta t}(\Delta t_i)$ has never been investigated for a particular band limited source signal. But the amplitude probability density function and first derivative probability functions have been investigated[6,29]. Therefore, assuming we know the first derivative density function for a particular source signal, we can find $p_{\Delta t}(\Delta t_i)$.

Δt_i corresponds to the time interval between two consequent

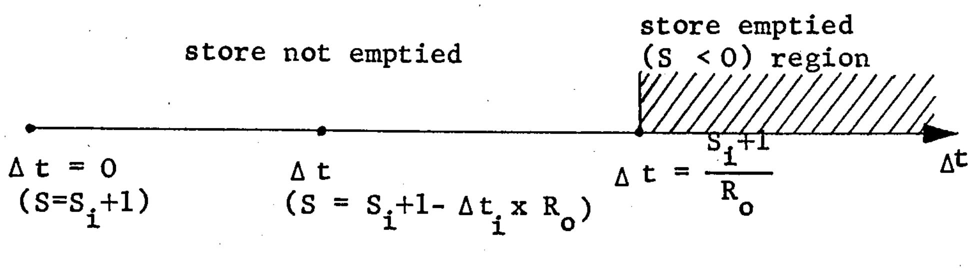


Fig.4.1 INTEGRATION REGIONS OF $p(\Delta t)$.

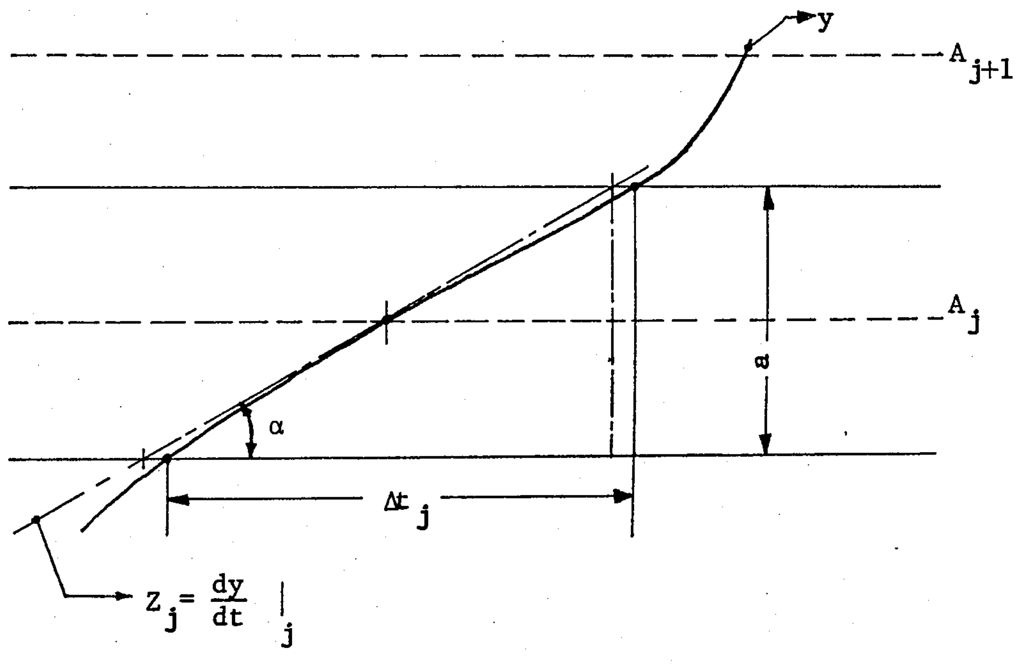


Fig.4.2 CONSECUTIVE AMPLITUDE LEVEL CROSSINGS.

amplitude level crossings (see Fig.4.2). For large values of m, $a/\Delta t_i$ can be accepted approximately equal to the derivative of the source signal waveform at this point.

If we know $p_z(\frac{dy}{dt})$, the probability density function of first derivative, $p_{\Delta t}(\Delta t_i)$ can be found. Putting $z = \frac{dy}{dt}$, and accepting that there is no maximum or minimum within Δt_i period.

$$p_z(z).dz = p_{\Delta t}(\Delta t_i).d(\Delta t_i) \quad \text{for} \quad |z| = \frac{a}{\Delta t_i} \qquad 4.12$$

Therefore

$$p_{\Delta t}(\Delta t_i) = p_z(\frac{a}{\Delta t_i}).\frac{a}{\Delta t_i^2} \qquad 4.13$$

can be written.

By substituting $p_{\Delta t}(\Delta t_i)$, and putting $\Delta t_i = \frac{S_i+1-S}{R_o}$ to equation 4.11 :

$$p(S) = +\int_o^{\infty} \frac{a \times R_o}{(S_i+1-S)^2} \; p_z(\frac{a \times R_o}{S_i+1-S}) \; . \; p(S_i).dS_i \qquad 4.14$$

The solution to this integral equation gives the probability density function of the store occupancy for $0 \leq S < \infty$. It must be noted, however, this equation is valid considering that calculations are done in the active speech region where $\frac{dy}{dt} \neq 0$. Long silence gaps must be considered separately. In this case $p(\Delta t)$ can be accepted as an exponential distribution as 'a level change' will become a rare event. (See Appendix).

4.3.2 Overflow and Underflow of the Store

Since the store occupancy density function is known, to find the probability of the store occupancy being larger than S_{max} we calculate the area under the probability curve above S_{max},

$$P(S > S_{max}) = \int_{S_{max}}^{\infty} p(S).dS. \qquad 4.15$$

This gives us the stationary overflow probability function, but does not provide us with a criterion to judge the subjective quality of the system. Because, although $P(S > S_{max})$ may be the same for two different signals of the same system, subjective quality test results of the reconstituted signals may differ. For both cases, the overflow condition will last for $1/R_o$ seconds. But, the signal which has the higher rate of change at the period of overflow would be affected more than the signal having less rate of change at the same period. As a result, there will be a subjective quality difference between the reconstituted forms of these signals.

If the signal consists of voltage spurts having high rates of change, the loss of pulse groups will be more than when the rate of change of the voltage is low, although the overflow period will be equal in both cases. Therefore, it must be clear that there are two different affects of overflow;

(1) The lost time because of overflow,

(2) The loss of information because of overflow.

So the signal voltage having high rates of change will be more affected as more information is lost about it.

It is obvious that distortion caused by overflow is closely related to the statistical properties of the signal voltage.

When the overflow occurs, the transmitter store does not register any more information. Hence the unit quantum changes in the amplitude of the signal voltage would not be reconstituted at the receiver. (Fig.4.3). A feature of overflow distortion is a d.c. shift in the reconstituted signal voltage. The amplitude of this d.c. shift varies, depending on the number of pulse groups

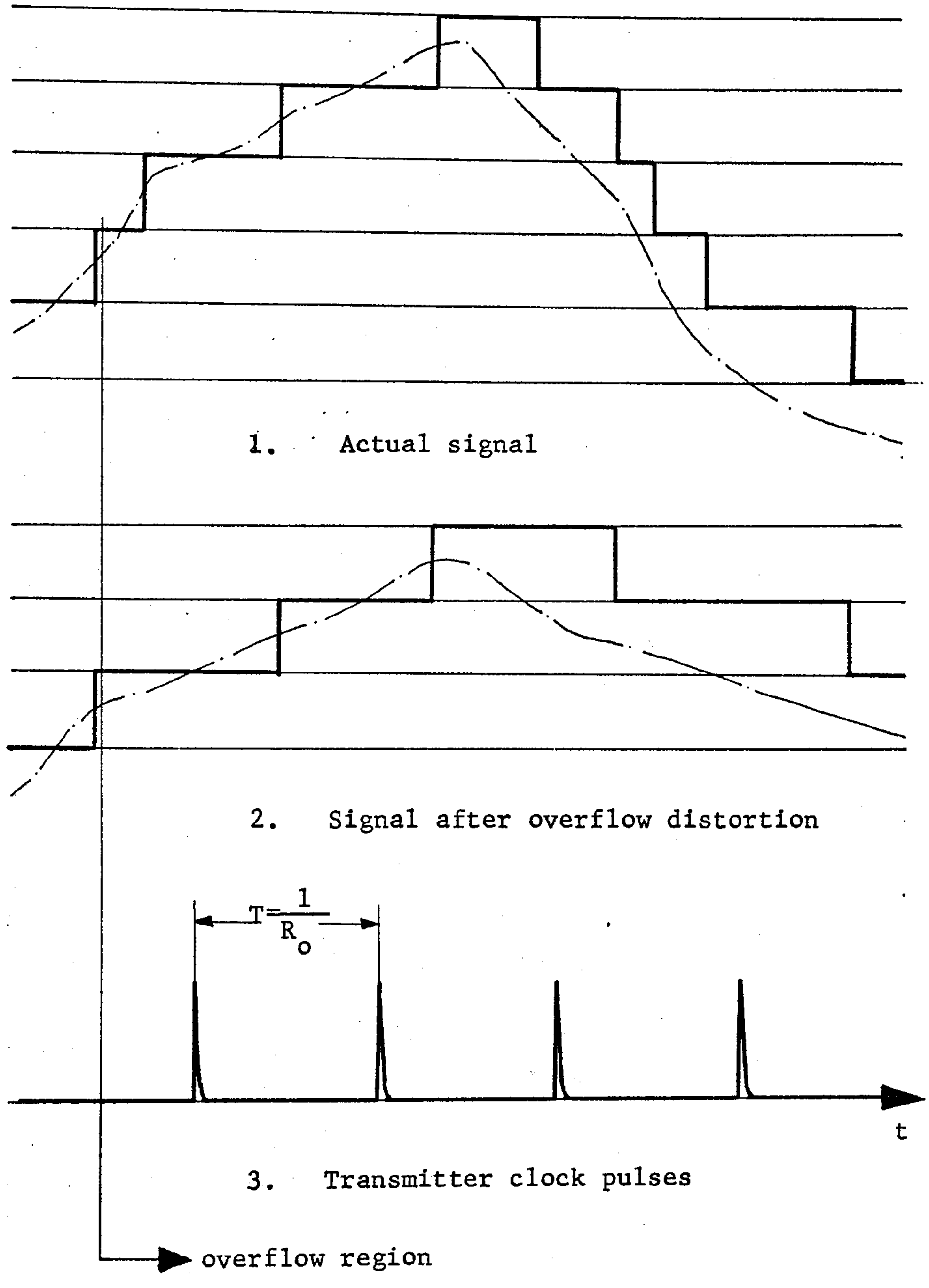


Fig. 4.3 OVERFLOW DISTORTION

which have been missed as the overflow period continues. For the speech signals this distortion effect appears as a 'click' noise in the practical observations.

We can find the maximum possible d.c. shift of the reconstituted voltage considering that the source signal is a band limited signal. The maximum rate of change will occur with a signal voltage having the maximum slope. If we consider the number of level changes of the signal voltage as N_x, in an overflow period

$$\frac{dy}{dt} = \frac{a \times N_x}{1/R_o} \text{ volts/sec.} \qquad 4.16a$$

To find the maximum rate of change, the maximum value of the slope has to be calculated:

$$N_{x\ max} = \frac{1}{a \times R_o} \times \left.\frac{dy}{dt}\right|_{max} \qquad 4.16b$$

As we are dealing with stationary random processes $\frac{dy}{dt}$ is not always known. We can only know the amplitude probability density function or first derivative probability density function of the amplitude of such signal voltages. By using 4.16a we can find $p(N_x)$, the probability density function of the number of the missed levels.

We define the overflow distortion as the expected value of the number of level changes in an overflow period. As we know the occupancy density function,we can write the conditional level change probability function in an overflow period as

$$p\left[N_x/S > S_{max}\right] = p(N_x) \times P\left[S > S_{max}\right] \qquad 4.17$$

subject that N_x and S are two independent variables. It is clear that there is no relation whatsoever between the rate of amplitude change and store occupancy, so we can write 4.17.

If we substitute $p(S > S_{max})$ by equation 4.15 and find $P(N_x)$, the conditional level change probability function would be found.

$P(N_x)$ can be found by using Poisson's distribution formula assuming that the outcome of the event of level changes are equally likely for the overflow period[2,7].

To find the overflow distortion we can write,

$$d_{of} = E\{N_x\} = \int_{o}^{N_{x\,max}} N_x . p\left[N_x / S > S_{max}\right] . dN_x \qquad 4.18$$

N_{max} is known for a band limited signal by using 4.16b. Therefore calculation of d_{of} would be a straightforward process. Results are applied to the speech waveform as the source signal, in Chapter 8.

In many ways, underflow is not so important as overflow. It is only important as far as the redundancy is concerned. If the store underflows the transmitted pulse groups will not carry any useful information. Therefore, these groups will occupy the transmission channel unwantedly. To minimise the flow of redundant information through the transmission channel, both store size and transmission frequency have to be chosen after careful consideration. Underflow will occur less frequently as the store size is increased and the transmission rate decreased.

If the transmission frequency differs from the average input rate, there will be a gradual increase or decrease in occupancy of the store. This will lead to a corresponding increase in the amount of overflow or underflow. This is shown in the graphs of Figs.8.1, 8.2 and 8.3. Low values of transmission frequency result in large values of overflow, while the underflow is small; for high transmission frequencies the reverse is the case.

It is difficult to determine an optimum transmission frequency. The effects on the received signal caused by overflow and underflow are different: overflow represents lost information, while underflow represents unused transmission capability. The choice of the optimum transmission frequency must await the results of overall tests on the equipment, in which an acceptable level of distortion may be determined. However, for the purposes of this research some criterion for this choice must be established, and the transmission frequency has therefore been chosen to be that value at which the curves in Figs. 8.1, 8.2 and 8.3 cross each other.

Referring to Fig.4.2 and equation 4.6, when $\Delta t_i \geqslant \frac{1+S_i}{R_o}$ the store underflows. Therefore the probability of zero occupancy can be found by using this relation :

$$P(\text{underflow}) = P\left[\Delta t_i \geqslant \frac{1+S_i}{R_o} / S_i\right] \qquad 4.19$$

Obviously S_i and Δt_i are independent variables, therefore ;

$$P(\text{underflow}) = \int_0^\infty P(S_i)\, P\left[\Delta t_i \geqslant \frac{1+S_i}{R_o}\right] dS_i \qquad 4.20$$

gives the probability of underflow.

As the average input rate is dictated by the signal, with a changing transmission rate we can observe how the overflow distortion changes. Ideally, if the transmission rate is equal to the average input rate, the store will never be filled or emptied.

Mathematically, the optimum transmission rate has been calculated and given in Chapter 6, equation 6.8.

4.4 EFFECTS OF THE NUMBER OF AMPLITUDE QUANTA m ON THE STORE SIZE NEEDED

Although the optimum transmission rate, as defined above would be the same for different store sizes, if we reduce the number of amplitude quanta levels, the average input data rate to the store will be decreased. As the levels are not so close together, time interval measurements will be made less frequently. This will naturally mean a reduction in the transmission rate. This will also permit us to use smaller store sizes. A reduction in store size will aid economy, which is rather important for practical systems.

However, it is not possible to make a drastic reduction in the number of the amplitude levels, because the coarser the amplitude quanta, the lower the signal to quantizing noise ratio becomes. Therefore this limitation draws yet another boundary for the choice of the system parameters.

The effect of m on the signal to quantizing noise ratio is investigated in Chapter 5.

CHAPTER 5

NOISE CHARACTERISTICS OF THE SYSTEM

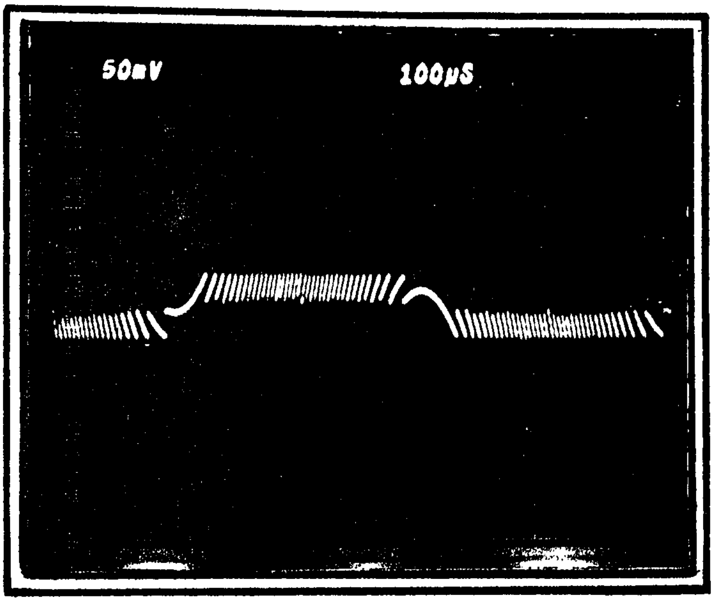
CHAPTER 5

5.1 NOISE CHARACTERISTICS OF THE SYSTEM

It is explained in previous chapters that the crossing of the signal waveform from the predetermined amplitude levels is being detected by the level change detector. Therefore, the time interval measured is between two consecutive amplitude quantum levels. We can see from this explanation that a quantizing effect takes place in this type of modulation, and of course, causes some impairment to the signal.[5] A similar problem occurs in PCM systems , namely quantizing noise. If it had been possible to make an accurate measurement of the time intervals and transmit this information in an analogue form of signal, time quantizing effect would not be accounted for. At this stage we will investigate only the effect of the amplitude-quantization.

Quantizing noise is not like random noise. It is correlated to the signal which is being quantized. Therefore, subjective assessments would be different and final judgement for the quality definition should be carried out by subjective tests. The quantization effect can be seen from the pictures taken from a CRO, when a sinusoidal signal voltage is applied to the input of a quantizer. In this application quantum levels are selected with equal intervals. (See picture 5.1).

If the signal voltage range is divided into m quantized amplitude levels, y_j will represent all the voltage values between $y_j - a/2$ and $y_j + a/2$. If we know the probability density of the amplitude distribution, it can be accepted that for one quantum



Photograph 5.1 QUANTIZING NOISE WAVEFORM

Source Signal	Sinusoidal voltage	Random noise voltage
Signal level	0 dB	0 dB
Noise level	- 39.2 dB	- 36.3 dB

Table 5.1

QUANTIZING NOISE MEASUREMENTS FOR 128 QUANTUM LEVELS.

interval, as it is small enough, this distribution does not change. Therefore, for this interval $p(y) = p(y_j)$, can be written. The squared error caused by the quantizing can be written as follows:

$$e_j^2 = \int_{y_j - a/2}^{y_j + a/2} (y - y_j)^2 . p(y) . dy \simeq \frac{1}{12} p(y_j) \ a^3 \qquad 5.1$$

As the quantum is very small with respect to the total range $p(y_j) \times a$ can be accepted as the probability of the signal being in the jth level. Each quantized level will add another noise component, so that squared error will consist of the sum of e_j^2's.

$$e^2 = \sum_j e_j^2 = \frac{1}{12} \ a^2 \qquad 5.2$$

From the picture (5.1), it is possible to see the waveform of the error signal. The above calculation gives the same result as the error signal's mean-square value, assuming it as a triangular waveform with peak to peak (a) volts deviation.

This noise voltage have a very wide frequency spectrum. To understand the fraction of this spectrum which overlaps with the actual signal spectrum, we have to find the spectral distribution of the quantizing noise.

In PCM systems sampling process, (amplitude sampling), results in the folding back of the spectrum, causing an increase in the noise level within the signal bandwidth. (Fig. 5.1). As we are not introducing a sampling process with VTSIP only a small fraction of the noise energy will be transmitted. This is obviously an advantage of the VTSIP system over the PCM.

In the VTSIP system, the noise as the result of the time quantizing process, will be much less effective than the noise

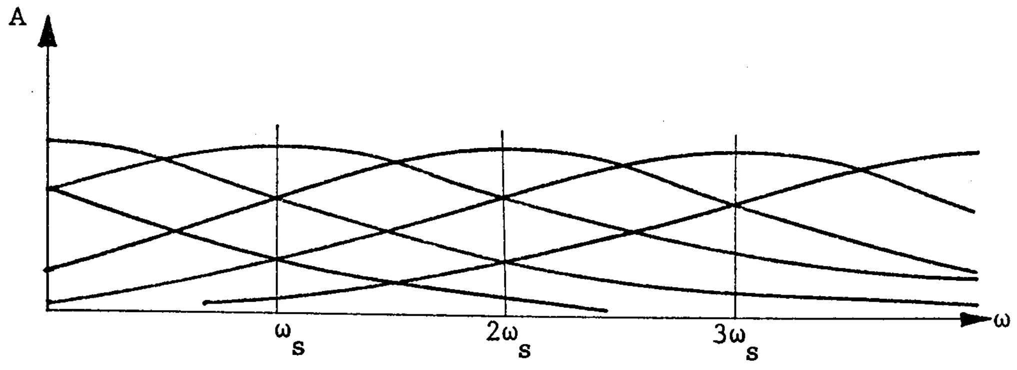


Fig. 5.1 QUANTIZING NOISE SPECTRUM AFTER SAMPLING.

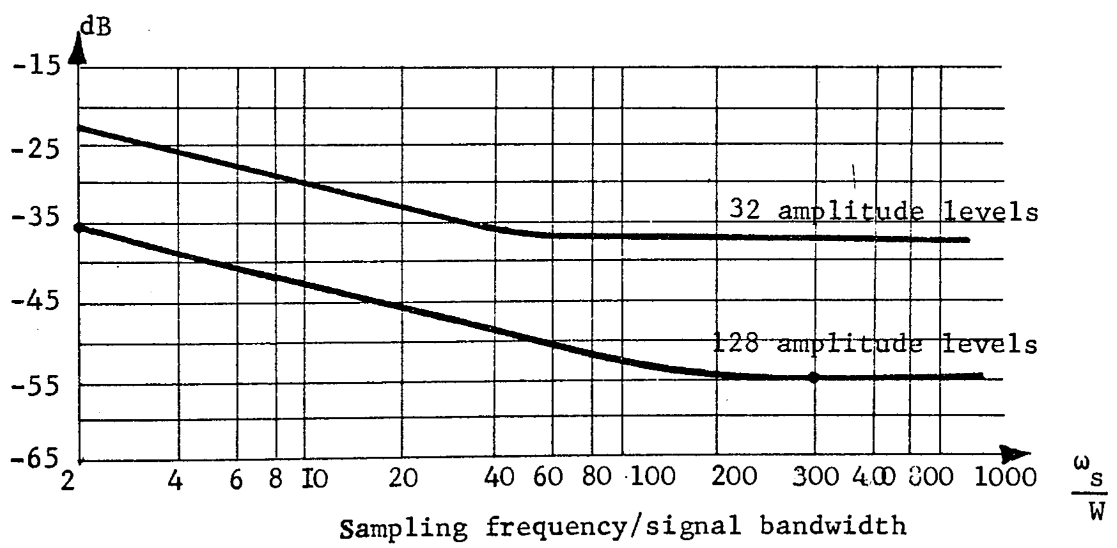


Fig. 5.2 QUANTIZING NOISE LEVEL BELOW SIGNAL INPUT POWER.
(this curve is taken from (4))

components as the result of the coarser time quantizing in PCM systems. As we are quantizing time-interval measurements between adjacent quanta levels with a very small unit time quantum, this is of course expected. Therefore, to establish a criterion for the noise specifications of the system, a series of subjective tests has to be carried out in addition to the theoretical calculations. This is not as easy as the above calculation, but somewhat similar. This will be dealt with in section 5.2.

As far as the calculation of the noise spectrum is concerned, it is again similar to the calculations which have been done previously by several investigators[24,26,27]. This time sampling process will not be considered. However, time quantization effects will be investigated. This is also dealt with in the following section.

A noise spectrum of this kind is quite wide. For a band-limited source signal, its quantization noise also is band-limited as they have a correlation between each other. Electrically the noise voltage will be a series of triangular voltages having random slopes. The bigger the number of quantum levels the smaller the noise amplitude will be. Therefore the length of each triangle will be proportional to the signal bandwidth, the signal amplitude and the number of quantum levels.

Prior to the calculation of the noise spectrum, it is in fact necessary to define the signal spectrum. The signal voltage consists of the ensemble of various signals having different frequencies and amplitudes. By investigating the power spectrum we can make some assessment of the relative magnitudes of the different frequency components of the actual signal. So we learn some useful information about the statistical properties of the signal, as we cannot define it analytically. For the stationary random signals we can define the power spectrum by using the auto-correlation function .

$$W(T) = \lim_{t \to \infty} \frac{1}{t} \int_{-t/2}^{+t/2} y(x).y(x + T).dx \qquad 5.3$$

This was first proved by Wiener[30] , which of course has been the only way to calculate the power spectrum of a random signal. So far as the quantizing noise is concerned the same method can be used. But for the values of T which are other than 0 the noise signal can be accepted as uncorrelated. Only for when T = 0 has the correlation function a value. From equation 5.3, if T is made zero W(T) becomes a delta function, which means it has a uniform power spectrum.

$$W(0) = \lim_{t \to \infty} \frac{1}{t} \int_{-t/2}^{t/2} y(x)^2 \, dx \qquad 5.4$$

$W(T) = \delta(t)$ which corresponds

$F(\omega)^2 = 1$

This result is of course correct for random noise signals. But in the case of quantizing noise, $F(\omega)$ is a low-pass function rathher than a uniform function. Because of the correlation with the signal which is being quantized, the quantization noise spectrum has a cut-off frequency of ω_Q.

If we normalize the function which represents the quantizing error spectrum, so that its integration over the spectra is one, we obtain :

$$\int_0^\infty F(x) \, dx = 1 \qquad 5.5$$

As F(x) is in the form of a low pass function it can be written as:

$$F(x) = F(\omega/\omega_Q) \qquad 5.6$$

where $\omega_Q = A.\omega_h.m.C.$ $\quad C = \text{constant}.$

This constant must be defined. Some authors have defined this constant for quantizing noise[4]. We do not intend to discuss the evaluation of this function any further. By combining this result with equation 5.5 we can write,

$$E^2(\omega) = \frac{F(\omega/\omega_q)}{3m^2\omega_q} \qquad 5.7$$

as proceeded in (5),

where $\omega_Q = \pi\sqrt{\frac{2}{3}}\ \omega_h.m.A$ as found in (4).

To find the noise power corresponding to the signal spectrum we must integrate E^2 over the signal bandwidth.

If the calculation is done, noise power within the signal bandwidth becomes very small as it is spread over a very wide frequency spectrum.

Therefore, as we are going to use a low pass filter at the receiver when we are reconstituting the signal voltage, only a small fraction of the noise power will interfere with the actual signal.

However, as the time quantizing process takes place, the noise power will increase a few dBs. Because of the sampling with regular time intervals in PCM systems, the noise power which falls within the actual signal spectrum is much higher than the noise which is contributed by the time interval quantization. Obviously, a reduction in the number of quantized amplitude levels will result in an increase in the quantizing noise power. By adjusting the number of these levels it is possible to arrange the noise power within the signal spectrum so that it is equal to the noise power of the PCM system using more amplitude quantizing levels. This is clearly an advantage of the proposed system over PCM systems. If we make an analogy and consider the time interval quantization as a result of sampling with a period equal to the unit time interval quantum, ω_s/ω_h becomes a very large number, where ω_s is the sampling radial frequency. With PCM ω_s is usually chosen as $2\omega_h$, therefore as $\omega_s = 2\pi \frac{1}{\tau}$ and $\tau << \frac{2\pi}{\omega} \, 2\omega_h$ in our case $\omega_s >> 2\omega_h$. If we refer to the curves drawn in Fig.5.2. it is obvious that for the same number of levels the signal to noise ratio obtained in VTSIP is much better than in PCM.

If m is large enough equation 3.3 can be written as:

$$\Delta t_{min} = \frac{1}{\pi m f_{max}} \qquad 5.8$$

Assuming $\tau = \Delta t_{min}$ is selected, therefore $\omega_s = \frac{2\pi}{\tau}$ and $\omega_h = 2\pi f_{max}$,

$$\frac{\omega_s}{\omega_h} = \pi . m \qquad 5.9$$

For m = 128 levels from the curve (Fig.5.2) the signal to noise ratio becomes 55 dB, whereas for the same number of levels and for the $2\omega_h$ sampling frequency it is only 35 dB for a PCM system. From the curve it is evident that the same signal to noise ratio can be obtained by the VTSIP system using only 32 quantum levels. (These curves have been taken from(4), Fig.5.2).

As ω_s/ω_h is very large for the time interval quantizing, there is not much difference between the time interval quantizing and unsampled quantizing. Therefore the quantized signal can be treated as unsampled. The spectrum of this type of quantizing noise depends very much on the actual form of the signal. With this research only the linear quantizing for both the time interval and amplitude is investigated. Remaining work for the search of the optimum quantizing processes still has to be completed. Discussion on this matter is given in chapter 10.

5.2 QUANTIZING NOISE

In section 5.1 quantizing noise as the result of only amplitude quantizing has been calculated. With our system an extra noise component is added because of the time-quantizing, to that calculated. Considering the characteristics of the level change detector it is possible to calculate the added quantizing noise.

Initially, after the first level crossing is detected, the counter in the transmitter starts to count the length of the time interval until the next level is crossed. The intervals are measured in multiples of τ which is the unit time quantum. Therefore, if the signal voltage crosses the next amplitude level before the completion of the period τ, the level change will not be detected until the end of this period. When the signal voltage is reconstituted in the receiver, as the level changes will not be precisely positioned in time a corresponding amplitude error will occur. For short time intervals the time displacement will be proportionally larger than when the time interval is longer.

The event of level crossing occurs as in Fig.5.3. Assuming that the level A_j represents the signal voltage values between the lower threshold of this level and the upper threshold, the squared error can be found as follows:

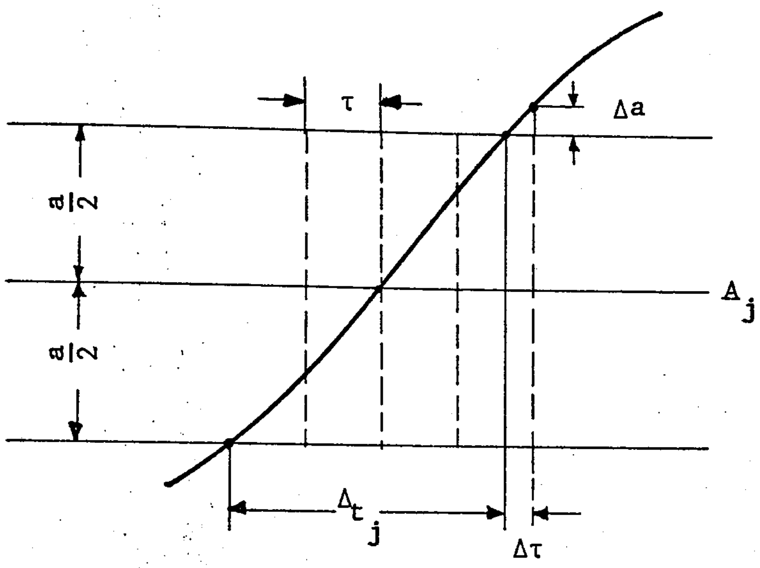


Fig. 5.3 EFFECTS OF TIME-INTERVAL QUANTIZING.

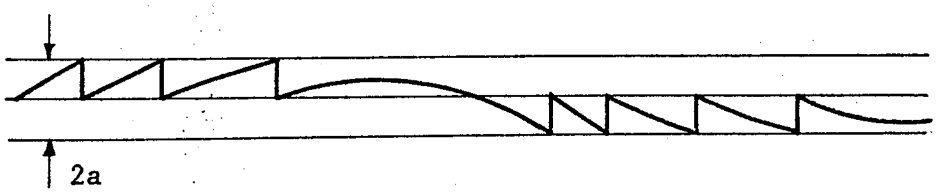


Fig. 5.4 QUANTIZING NOISE WAVEFORM

$$e_j^{\,2} = \int_{A_j-a/2+\Delta a_{av}}^{A_j+a/2+\Delta a_{av}} (A-A_j)^2 . p(A) . dA \qquad 5.10$$

where p(A) is the probability density function and Δa_{av} is the average displacement of amplitude because of the time interval quantizing. The calculation is similar to that in section 5.1, except the addition of Δa_{av} at the upper boundary.

For every amplitude level, depending on the signal slope the value of Δ a changes. As we are dealing with band limited random processes the average Δa_{av} can be found.

Bu substituing $\varepsilon = A - A_j$, $d\varepsilon = dA$ in equation 5.10, and assuming that for this small interval P(A) is a uniform distribution function, and $p(A) = p(A_j)$ in this interval,

$$e_j^{\,2} = \int_{-a/2+\Delta a_{av}}^{a/2+\Delta a_{av}} \varepsilon^2 . p(A_j) . d\varepsilon \qquad 5.11$$

$$e_j^{\,2} = p(A_j) . \left. \frac{\varepsilon^3}{3} \right|_{-\frac{a}{2}+\Delta a_{av}}^{+\frac{a}{2}+\Delta a_{av}} = \frac{1}{3} p(A_j) \left[(\frac{a}{2} + \Delta a_{av})^3 - (\Delta a_{av} - \frac{a}{2})^3 \right]$$

$$= \frac{P(A_j)}{3} \left[\frac{a^3}{4} + 3a \;\; \Delta a_{av}^2 \right]$$

From the previous section we know that only the amplitude quantizing squared error is :

$$e^2 = \frac{1}{12} \; a^2$$

To find the total error because of the amplitude and time quantizing, we have to add the errors contributed by each amplitude quanta :

$$e_Q^2 = \sum_j e_j^2 = \sum_j \frac{P(A_j)}{12} a^3 + \sum_j P(A_j)a.\Delta a^2_{av}$$

As the amplitude probability distribution uniform in this small interval, $[p(A_j).a] = p_j$, and $\sum_j P_j = 1$, therefore,

$$e_q^2 = e^2 + \Delta a^2_{av} \qquad 5.12$$

If τ is chosen smaller than any likely time interval we can write:

$$\frac{\Delta a_j}{\Delta\tau_j} \simeq \frac{dy}{dt}\bigg|_{t=t_j} = z_j \qquad 5.13$$

a_j is of course a random variable, therefore to find Δa_{av} we have to calculate its expected value,

$$\Delta a_{av} = E\{\Delta a_j\} = \int_o^{\Delta a_{max}} \Delta a_j \, p(\Delta a_j) \, d(\Delta a_j) \qquad 5.14$$

From 5.13, Δa_j is a function of $\Delta\tau$ and the first derivative of the signal. Therefore, as known from probability theory

$$E\{\Delta a_j\} = E\{\Delta\tau . z\} \qquad 5.15$$

$\Delta\tau$ and z can be accepted as independent variables for such a process. So, as known from probability theory:

$$E\{\Delta a_j\} = E\{\Delta\tau\}.E(z) \qquad 5.16$$

To find the expected values of both $\Delta\tau$ and z we must know $p(\Delta\tau)$ and p(z). For speech signals p(z) has been investigated[29] But in the following calculations E(z) is left in integral form.

$$E\{\Delta a_j\} = \int_o^{\tau} \Delta\tau p(\Delta\tau) d(\Delta\tau) . \int_o^{z_{max}} z.p(z).dz \qquad 5.17$$

$\Delta\tau$ is uniformly distributed between 0 and τ. Therefore $p(\Delta\tau)$ is equal to $\frac{1}{\tau}$. The expected value of $\Delta\tau$ is then,

$$E(\Delta\tau) = \frac{1}{\tau} \int_o^{\tau} \Delta\tau \, d(\Delta\tau) = \frac{\tau}{2} \qquad 5.18$$

By substituting $E\ \{\Delta\tau\}$ in 5.17,

$$E\ \{\Delta\ a_j\} = \frac{\tau}{2} \int_o^{z\ max} z.p(z).dz \qquad 5.19$$

Where z_{max} is the maximum value of the slope of the signal voltage. If the highest frequency in the signal band is f_h maximum slope for this band limited signal voltage will be $z_{max} = f_h.\pi.a.m.$ Referring to equations 5.12 and 5.13 the squared error is,

$$e_q^{\ 2} = \frac{1}{12}\ a^2 + \left[\frac{\tau}{2} \int_o^{f_h \pi.a.m} z.p(z)^{dz}\right]^2 \qquad 5.20$$

The additional term of this expression is because of the time quantizing which is implemented in the proposed system. If we can find p(z), which is the probability density function of the first derivative of the signal, we can calculate this additional noise component.

After the reconstitution of the signal voltage at the receiver, only the noise components which are within the signal bandwidth will affect the signal to noise ratio. As m is chosen large enough the noise spectrum will be very wide and therefore a big proportion of the noise energy will be out of the signal band. Therefore, the signal to noise ratio will be better than expected. If we find the error signal (quantizing noise) spectrum we can calculate the noise power within the signal band.

If we look at the quantizing error waveform which is shown in Fig.5.4, it consists of triangular segments having different slopes. The average segment length is equal to the expected value of Δt_j.

The signal voltage is expressed with a monothonic increasing or decreasing function between two adjacent quantum levels. Therefore, it can be approximated by a linear segment connecting the two adjacent crossing points of the signal voltage with the quantum levels, so that we can write :

$$\Delta t_j = a \frac{1}{\left|\frac{dy}{dt}\right|_j} \qquad 5.21$$

It is knownthat since we know the probability distribution of the first derivative of the signal we can find $p(\Delta t_j)$. From $p(\Delta t_j)\, d(\Delta t_j) = p(y'_j)$, it follows that :

$$p(\Delta t_j) = \frac{a}{\Delta t_j^2}\, p_{y'_j}\left(\frac{a}{\Delta t_j}\right) \qquad 5.22$$

To find the average Δt_j we have to calculate its expected value. Therefore,

$$\Delta t_{av} = E\{\Delta t_j\} = \int_{\Delta t_{j\,min}}^{\Delta t_{j\,max}} \Delta t_j\, p(\Delta t_j)\, d(\Delta t_j)$$

$$= \int_{\frac{1}{2\pi f_h} \arcsin \frac{2}{m}}^{\frac{1}{f_1}} \frac{a}{\Delta t_j}\, p_{y'_j}\left(\frac{a}{\Delta t_j}\right) d(\Delta t_j) \qquad 5.23$$

Obviously for a stationary random signal, the expected value of Δt_j gives us some information about the average occurrence of data.

Although the segment length is equal to $E\{\Delta t_j\}$ in average, there are times when segments are much shorter than this value. But it is evident that these segment lengths are in multiples of τ which is the unit time quantum. Therefore, it can be said that the error voltage spectrum has a crossover point at a frequency $f_c = \frac{k}{\tau}$, assuming the spectrum is a low pass function of frequency. If we call this function $g(\omega/\omega_c)$, and integrate it over the spectrum the result will be equal to the squared error function which gives the noise power. The detailed calculations are rather complicated. Some investigators[4,27] have found the power spectrum expressions for the amplitude quantizing error. Bennett[4] also derived the formulae for the spectrum of the error signals after being quantized in time as well as amplitude.

By using the Wiener's correlation theorem and Fourier cosinus theorem Bennett has found that the spectrum of the amplitude quantizing error is :

$$F_o\left(\frac{f}{f_h}\right) = \frac{1}{2\pi^{\frac{7}{2}} 4^{\frac{3}{2}(m-3)}} \sqrt{\frac{3}{2}} \sum_{n=1} \frac{1}{n^3} e^{-\frac{3}{8\pi^2 4^{m-3}}\left(\frac{f}{f_h}\right)^2} \qquad 5.24$$

For the sampling case Bennett sums up all contributions from each harmonic of the sampling rate beating with noise spectrum of quantizing error,

$$G_f\left(\frac{f}{f_h}\right) = F\left(\frac{f}{f_h}\right) + \sum_{n=1} \left[\left(F_{n(f_s-f)} + F_{n(f_s+f)}\right)\right] \text{where } 0 \leqslant f \leqslant f_{s/2}$$

To find the error signal within the signal bandwidth by substituting $\frac{f}{f_h} = 1$,

$$G_o = F_o(1) + \sum_{n=1} \left[F_o\left(n\frac{f_s}{f_h} + 1\right) + F_o\left(n\frac{f_s}{f_h} - 1\right)\right] \qquad 5.25$$

A set of curves calculated from equation 5.25 is plotted by Bennett in his paper. We have plotted these curves in Fig. 5.6. From these curves an immediate comparison of the signal to noise ratios of the PCM and VTSIP can be made easily.

From this result it is obvious that VTSIP system has a much better performance as far as signal to noise ratio only, is concerned. By reducing the time quantum unit τ it is possible to achieve much better signal to noise ratios. But,if the system does not require that by reducing the number of amplitude quantizing levels signal to noise ratio can be decreased. Depending on the quality wanted this can be decided after doing a series of subjective quality tests.

Some noise measurements have been made for the band limited noise after being quantized in the amplitude quantizer. The results are given in Table 5.1, both for sinusoidal voltage and random noise voltage as input signals and corresponding noise spectrums of these signals have been shown in photographs 5.2 and 5.3. The corresponding noise voltage waveform for the sinusoidal voltage, is shown in the photograph 5.1, which has been taken from the screen of a CRO. At the present stage, it was only possible to make measurements for the amplitude quantized signal waveform. It is suggested for further investigations that noise measurements be made for various amplitude quantum levels. So, it can be verified that the VTSIP system having lesser amplitude quantizing levels than the PCM system gives the same signal to noise ratio performance.

The circuits used for these measurement are shown in Fig.5.7. The filter band was arranged to coincide with that of speech. A gaussian noise source was used to create speech-like voltage waveforms. When the switch is closed to A the voltmeter and the CRO indicate the value of the wideband error voltage. But, when the switch is at B, the filter (F_2) passes only the

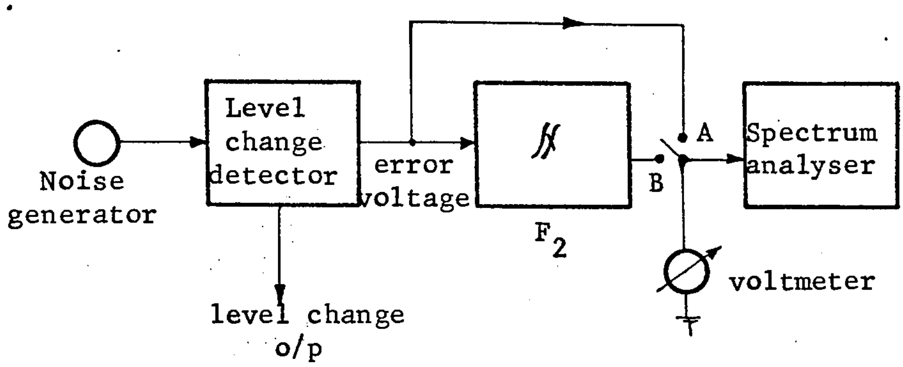


Fig. 5.5 NOISE MEASUREMENT METHOD.

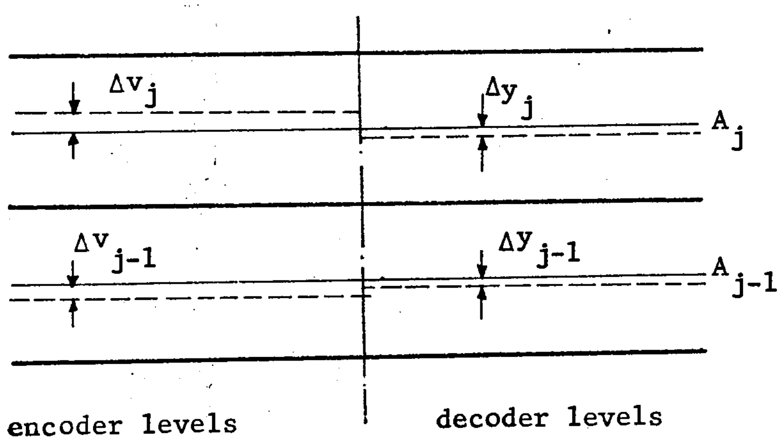
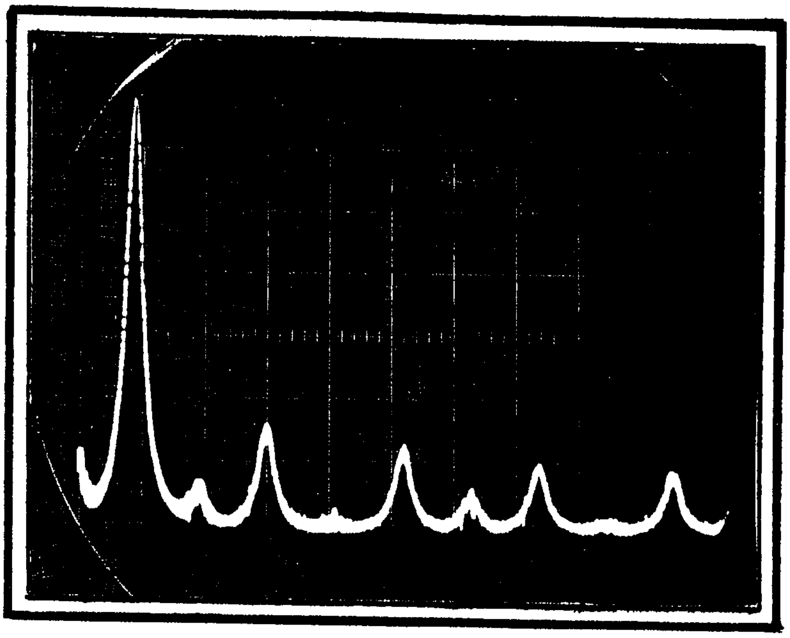
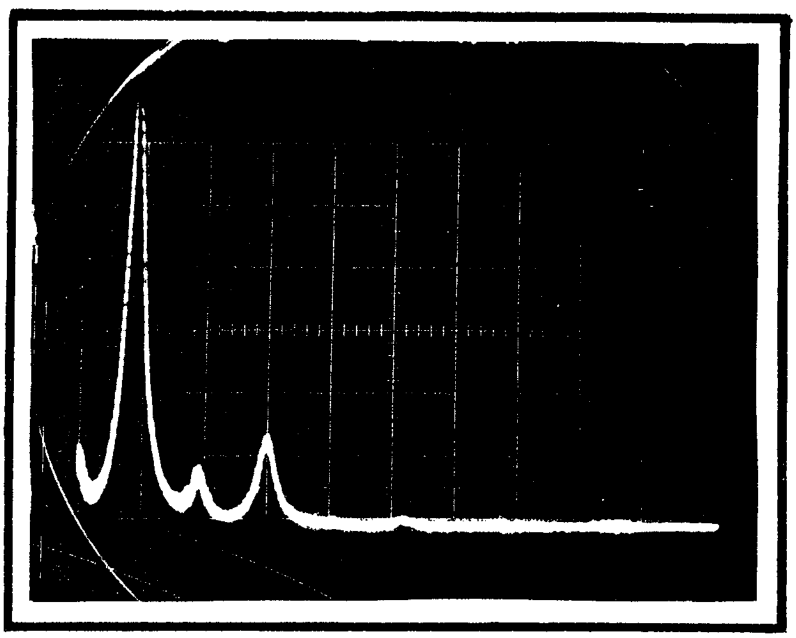


Fig. 5.6 IRREGULARITIES IN QUANTIZING LEVEL



(a)



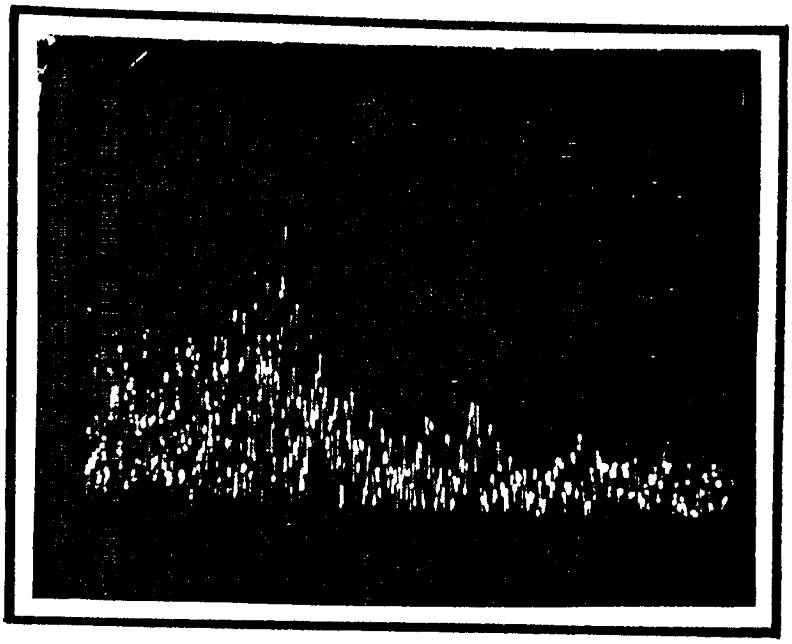
(b)

Photograph 5.2

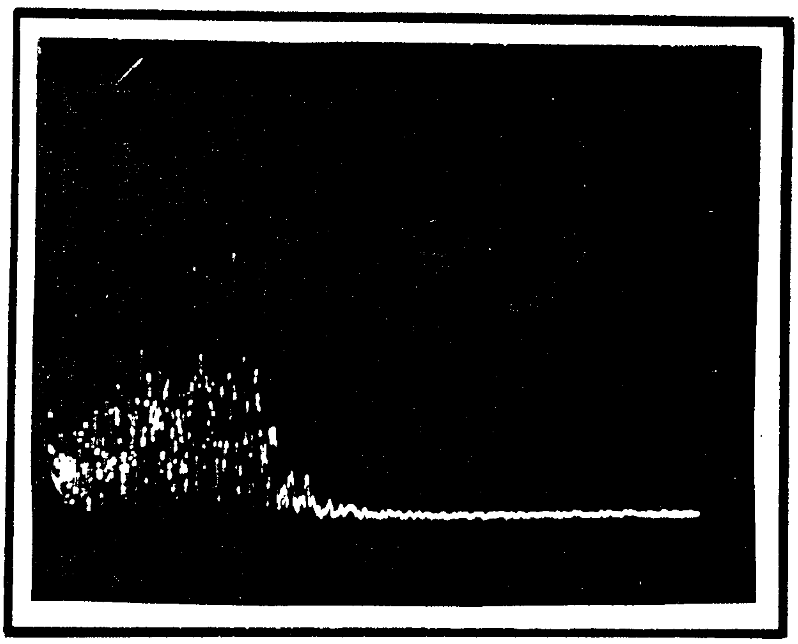
NOISE SPECTRUMS FOR SINUSOIDAL INPUT VOLTAGE

(a) Complete noise spectrum

(b) Noise spectrum after noise has been filtered.



(a)



(b) 1 cm = 1 kHz

Photograph 5.3

NOISE SPECTRUM FOR RANDOM NOISE VOLTAGE AS INPUT SIGNAL

(a) Complete noise spectrum

(b) Noise spectrum within the 3400 Hz band signal band, therefore the noise voltage, indicated by the voltmeter will be less than the previous voltage. The input voltage is arranged to have an rms value of ¼ of the peak clipping level, so that only for 1 per cent of time does the peak clipping occur. This is generally accepted by designers of the PCM systems.[5]

5.3 IRREGULARITIES IN QUANTIZING

In section 5.1 and 5.2 all calculations are carried out with the assumption that all the amplitude and time quanta are accurately placed. In practice, the quantum values may change randomly in time. The quantizer consists of electronic components having different characteristics within a tolerance. Over time, these characteristics may change resulting in the displacement of the time and amplitude quanta. It is obvious that we cannot construct ideal mechanisms doing the quantizing processes, because of these reasons. As we cannot predetermine these irregularities, it is also impossible to define the exact crossing point of the signal voltage with a quantum level. This means there is an uncertainty region around every quantum level. Similarly as the time quantum may change a displacement of time will occur. The irregularities of this type may result in the occurrence of an indecisive output.

In addition to the irregularities, because of the time varying characteristics of the electronic components in the quantizer, the random appearance of additional noise voltage on the signal will cause another type of indeterminancy. The analysis of the effects of these irregularities is very complex, and to derive objective conclusions from the result is difficult, especially after considering that similar irregularities will occur in the receiver when the received digital signal is being converted into analogue form.

In Fig. 5.6 the voltage levels and the irregular shifts of these levels are shown. If we add the change in the time interval quantum to this picture the obvious complexity of the process can be easily seen. The quantizer levels are shown in the left hand side of Fig.5.6, and the corresponding decoder levels are shown at the right hand side. For the set of $v_1 \ldots v_j \ldots v_m$ quantizer levels, a set of $y_1 \ldots y_j \ldots y_m$ decoder levels correspond.

We know from equation 5.7 that

$$e_j^2 = \int_{-\frac{a}{2} + \Delta a_{av}}^{\frac{a}{2} + \Delta a_{av}} \varepsilon^2 . p(A_j) . d\varepsilon \text{ where } \varepsilon = A - A_j$$

If the level displacements of the quantizer are $\Delta v_1 \ldots \Delta v_j \ldots \Delta v_m$ at time t, and the corresponding decoder level displacements are $\Delta y_1 . \Delta y_j .. \Delta y_m$ at time $t + t_d$ where t_d is the total system delay, new levels can be represented as follows:

$$v_j = j.a + \Delta v_j \quad \text{where } j = 1 \ldots m \qquad 5.26$$

The corresponding decoder levels will be :

$$y_j = (j - \tfrac{1}{2}).\ a + \Delta y_j \quad \text{where, } j = 1 \ldots m \qquad 5.27$$

As it is known from section 5.2, because of the time quantizing an extra error is already added to the amplitude quantizing error. Supposing that there is no displacement at the quantizer levels, mean squared error, referring to equation 5.11,can be written as :

$$e_j^2 = p(A_j) \int_{-\frac{a}{2} + \Delta a_{av} - \Delta y_j}^{+\frac{a}{2} + \Delta a_{av} - \Delta y_j} \varepsilon^2 . d\varepsilon$$

$$= p(A_j)\, \frac{1}{3} \left[\frac{a^3}{4} + 3a\, \Delta a^2_{av} + 3a\, \Delta y_j^2 - 6a\, \Delta y_j\, \Delta a_{av}\right]$$

$$= p_j\left[\frac{a^2}{12} + \Delta a^2_{av} + \Delta\, y_j^2 - 2\, \Delta y_j .\ \Delta a_{av}\right] \qquad 5.28$$

In addition to this, if we consider the level displacement at the quantizer, by replacing a by a' and Δy_j by $\Delta y'_j$, where

$$a' = a + \Delta v_j - \Delta v_{j-1} \quad \text{and}$$

$$\Delta y'_j = \Delta y_j - \frac{\Delta v_j - \Delta v_{j-1}}{2}\text{, so the squared error}$$

becomes:

$$e_j^2 = p(A_j)\, \left[a + \Delta v_j - \Delta v_{j-1}\right]\left[\frac{1}{12}\left(a + \Delta v_j - \Delta v_{j-1}\right)^2 + \left[\Delta a_{av} - \left(\Delta y_j - \frac{\Delta v_j - \Delta v_{j-1}}{2}\right)\right]^2\right] \qquad 5.29$$

The terms are obviously uncorrelated with each other in expression 5.29. As each quantizing level introduces an error component the total squared error is:

$$e^2 = \sum_j e_j^2$$

The term Δa_{av} also has to be modified considering the possible error in the time quantum. From equation 5.16 we know that Δa_{av} is the product of the expected value of both $\Delta\tau$ and z. As the frequency $1/\tau$ is produced by a reference oscillator, any shift in this oscillator frequency df is directly reflected by corresponding error in time quantum as $d\tau$. If we make this addition in the expression 5.16:

$$\Delta a_{av} = E\{\Delta a\} = \int_{o}^{\tau+d\tau} \Delta\tau \;\; p(\Delta\tau)\, d(\Delta\tau) . \int_{o}^{z_{max}} z.p(z).dz \qquad 5.30$$

All the parameters $d\tau$, Δy_j, Δv_j, Δy_{j-1}, Δv_{j-1} in these expressions are time dependent variables and their values change depending on the electronic component characteristics. Therefore, the design of the system must be made taking all the possible variations and their effects on the system specifications into consideration. It will be impracticable to carry this discussion any further as we are viewing the problem in a general way. However, for a rather restrictive investigation references [5] are available.

As previously discussed the random noise which is superimposed to the signal voltage is also a cause of error. As there is no correlation between the signal and the noise voltages whatsoever, simpler results can be obtained. These calculations have been done again by Professor Cattermole [5], for the case of amplitude quantizing. If we apply the same thoughts to our case we can find the effect of the input noise to the squared error.

If a small noise voltage v_n is added to the signal voltage, this will correspond to an increase in the squared error by v_n^2. Therefore the squared error will become :

$$e_j^2 = p_j'\left(\frac{1}{12} a^2 + v_n^2\right) \qquad 5.31$$

where, obviously p'_j will be changed. To find p'_j, we have to integrate p(A) between $(j-1)a - v_n$ and $(j.a - v_n)$, so that:

$$e^2 = \left(\frac{1}{12} a^2 + v_n^2\right) \sum_j \int_{(j-1)\, a.v_n}^{ja - v_n} p(A).dA \qquad 5.32$$

If we assume that $\sum p'_j = \sum p_j = 1$, which is usually an acceptable approximation,

$$e^2 = \frac{1}{12} a^2 + v_n^2 \qquad 5.33$$

can be written.

After this brief discussion about the effects of the imperfections of the amplitude and time interval quanta for uniformly quantized signals, it may be necessary to extend this discussion for non-uniform quantized signals.

CHAPTER 6

CHANNEL CAPACITY

CHAPTER 6

6. CHANNEL CAPACITY

In this section we will investigate the necessary channel capacity for the transmission of the information produced by our system. We can define the channel capacity as the maximum rate of information which can be conveyed through a communication channel with an acceptable error. If we are dealing with noiseless channels then instead of saying 'acceptable error' we can say 'errorless'.

Let us assume that we have a telecommunication channel which is designed for the conventional PCM system. Basically a channel like this has a capacity of $2f_h \log_2 m'$ bits/sec. If we want the VTSIP system to show an improvement over the PCM system, VTSIP signals must be conveyed from a channel having less capacity than the above mentioned channel.

Referring to equation 5.23, the average source information is produced with the $F = \frac{1}{\Delta t_{av}}$ occurrence frequency. If the average information produced by the system per message is $H(\Delta t_j)$, the average rate of information is

$$R = FH(\Delta t_j) \text{ bits/secs.} \qquad 6.1$$

where

$$H(\Delta t_j) = -\sum_j p(\Delta t_j) \log_2 p(\Delta t_j) \text{ bits/word} \qquad 6.2$$

As there is a one to one correspondence between the message and the code word, $H(\Delta t_j) = H(w_j)$ can be written, where w_j is the code word. $H(\Delta t_j)$ is always smaller or equal to $\log_2 2^l$ where l is the number of letters in the code word w_j. Assuming that every message Δt_j is coded into a word containing l_j letters:

$$-\sum_j p(\Delta t_j) \log_2 p(\Delta t_j) \leq \sum_j l_j \, p(\Delta t_j) \text{ can be written} \qquad 6.3$$

Therefore, if this equality is achieved a maximally efficient code can be designed. The condition for this is :

$$l_j = -\log_2 p(\Delta t_j) \text{ for all } j. \qquad 6.4$$

But by solving equation 6.4 :

$$p(\Delta t_j) = 2^{-l_j}$$

cannot be obtained every time, which is the necessary condition for most efficient coding. But by chosing l_j to the nearest integer, which makes this equality as close as possible, a very high efficiency code can always be designed. Redundancy is also an important problem of code designing. Therefore a minimum redundancy and maximum efficiency code must be designed to suit the system requirements.

A basic comparison with PCM can be made by calculating the information rate of the VTSIP system. As seen from equation 6.1, $R = F.H(\Delta t_j)$. From previous chapters we know $p(\Delta t_j)$, therefore,

$$R = \frac{-1}{t_{av_j}} \sum p(\Delta t_j) \log_2 p(\Delta t_j) \text{ can be written.}$$

Assuming that we have coded the message to a maximally efficient code,

$$H(\Delta t_j) = -\sum_j l_j \, p(\Delta t_j), \qquad 6.6$$

if we accept that $l_j = l_c$ for the maximum entropy

$$H(\Delta t_j) = -l_c \text{ is found} \qquad 6.7$$

By substituting $H(\Delta t_j)$ in equation 6.7,

$$R = +\frac{l_c}{\Delta t_{av}} \text{ bits/secs} \qquad 6.8$$

If channel capacity $C = 2f_h \log_2 m'$ is larger than R this is obviously the advantage of the system over PCM.

By substituting Δt_{av} from Equation 5.23,

$$R = \frac{1_c}{\displaystyle\int_{\frac{1}{2\pi f_h}\arcsin\frac{3}{m}}^{\frac{1}{2} f_1} \frac{a}{\Delta t_j^{\,2}} P_{y'_j}\left(\frac{a}{\Delta t_j}\right) d(\Delta t_j)} \qquad 6.9$$

is obtained.

If we find $\frac{C}{R}$, and if this ratio is larger than 1, that means our system needs less channel capacity than that needed for the PCM system

$$M = \frac{C}{R} = \frac{2f_h \log_2 m'}{\log_2 \dfrac{\pi \times f_h}{f_1 \arcsin \frac{2}{m}}} \int_{\frac{1}{2\pi f_h}\arcsin\frac{2}{m}}^{1/2\, f_1} \frac{a}{\Delta t_j^{\,2}} P_{y'_j}\left(\frac{a}{\Delta t_j}\right) d(\Delta t_j) \qquad 6.10$$

This ratio is defined as the merit factor of the VTSIP system. For individual systems this ratio will change. To find the advantage of the system over other pulse modulation systems this merit factor must be high enough, and at the same time the store size must be small enough. At first sight, the independency of M from the store size seems odd. But as the average occurrence of the information is not dependent on the store size, this result is not anomalous. As soon as we introduce the system errors then it can be seen that the store size is actually a limiting factor on the rate of distortion. By decreasing the store size, the overflow and underflow will increase and so the distortion will be higher. From equation 4.18 we know that $d_{of} = f(S_{max}, R_o)$. As R_o is dictated by equation 5.23 for the accepted optimum transmission rate criterion, d_{of} is only dependent upon S_{max}. S_{max} is therefore, the most important factor affecting the system overflow distortion. Similarly as the system delay will be:

$$\Delta_{VTSIP} = \frac{S_{max}}{R_o} \text{ secs} \qquad 6.11$$

When choosing S_{max} this delay also must be considered. There are objectionable limits for both Δ_{VTSIP} and d_{of} which must be found experimentally. Obviously this result will define for us an optimum store size to use.

CHAPTER 7

PROPERTIES OF THE SIGNAL TO BE TRANSMITTED

CHAPTER 7

7. PROPERTIES OF THE SIGNAL TO BE TRANSMITTED

In the VTSIP system the modulation process is carried out in a completely different way than for the other pulse modulation techniques. Therefore, it has been necessary to investigate the properties of the signal from this viewpoint. Of course, as we are dealing with random signals their statistical properties have to be defined. Without having this information, it will be an impossible task to design the proposed telecommunication system. Therefore, as we cannot define the signal analytically, a statistical approach to the properties of the signal has been made.

Several authors have investigated the properties of speech signals and other signals [6,20]. The aim of these investigators was to find the amplitude distributions, correlation functions, in other words the fundamental statistical properties of the signals. Only in Sciulli's paper[20] an approach to find the statistical characteristics of digital voice signals has been done. His aim was to find the necessary statistical characteristics of the speech signals for the data compression applications. In this research although we are trying to obtain a reduction in the transmission rate, the approach is completely different from the other investigators who try to compress PCM signals.

A most important characteristic of the signal which has to be known is the time interval distribution between two consecutive amplitude level crossings. This distribution is obtained by using the first derivative distribution of the signal. Obviously, signals with higher slopes cause an increase in the rate of information, and signals with low slopes cause a decrease in the rate of information. The average time interval or the expected value of the time interval is obviously very important. Variations from this average time interval cause excessive overflow and underflow. Therefore, besides time interval distribution its variance must be known.

The feasibility of the system is very much dependent on the signal characteristics. For example, if the signal consists of impulses with very high amplitudes and very short durations, the store will be filled in a very short time and, as it will take more time to empty the store, frequent overflow will occur. Especially, if the repetition frequency of the spurts of that kind increase, this will cause too much information loss. On the other hand, if the same signal contains very long silence gaps, this time, when the store is emptied, too much redundant information will be transmitted over the communication channel. Therefore, for some kinds of signals this method of modulation may not give profitable results.

First derivative distribution of a signal gives us the time-interval distribution as explained in section 5.2. By using the equation 6.2, we can find the entropy of the signal. According to the discussion in the above paragraph, the variation of the time interval from the average value must be limited for the particular signal which is being modulated. This will enable us to choose a transmission frequency which will not permit excessive overflow and underflow. By using these conditions we can define the statistical properties of the signal which are most suitable for the application of the VTSIP system. We know the average

Δt_i from equation 5.23. The entropy of the system is also known from equation 6.2. From the probability theory we know that the variance of a random variable is given by the formula

$$V(\Delta t_j) = (\Delta t_j^2)_{av} - (\Delta t_{j\,av})^2 \quad sec^2 \qquad 7.1$$

From equation 6.2 we know that the entropy is, (in continuous form)

$$H(\Delta t_j) = -\int_{t_{j\,min}}^{t_{j\,max}} p(\Delta tj)\log_2 p(\Delta t_j)\, d(\Delta t_j) \text{ bits per interval} \qquad 7.2$$

We can find the squared average value of Δt_j as:

$$(\Delta t_j^2)_{av} = \int_{t_{j\,min}}^{t_{j\,max}} \Delta t_j^2 \; p(\Delta t_j)\, d(\Delta t_j) \; sec^2 \qquad 7.3$$

The variations of the Δt_j from its mean value gives us an idea about the necessary store size to minimise the underflow and overflow. If the input rate to the store is equal to $\Delta t_{j\,av}$ a store will not be necessary. But, as the input rate changes, depending on the characteristics of the input signal, Δt_j will also change around the value of $\Delta t_{j\,av}$. If $p(\Delta t_j)$ is centred around $\Delta t_{j\,av}$ and the distribution curve has a shape as in Fig. 7.1, necessary store size will be less than when the curve has a flat form as shown in the same figure with dotted lines.

The signals having, for example, uniform time-interval distributions will be the most difficult to modulate. Because, Δt_j's can have any value equally likely between $\Delta t_{j\,min}$ and $\Delta t_{j\,max}$. That means the store size must be big enough to accommodate the amount of information which is likely to be produced with very high rates as well as low. In this case, R_o,

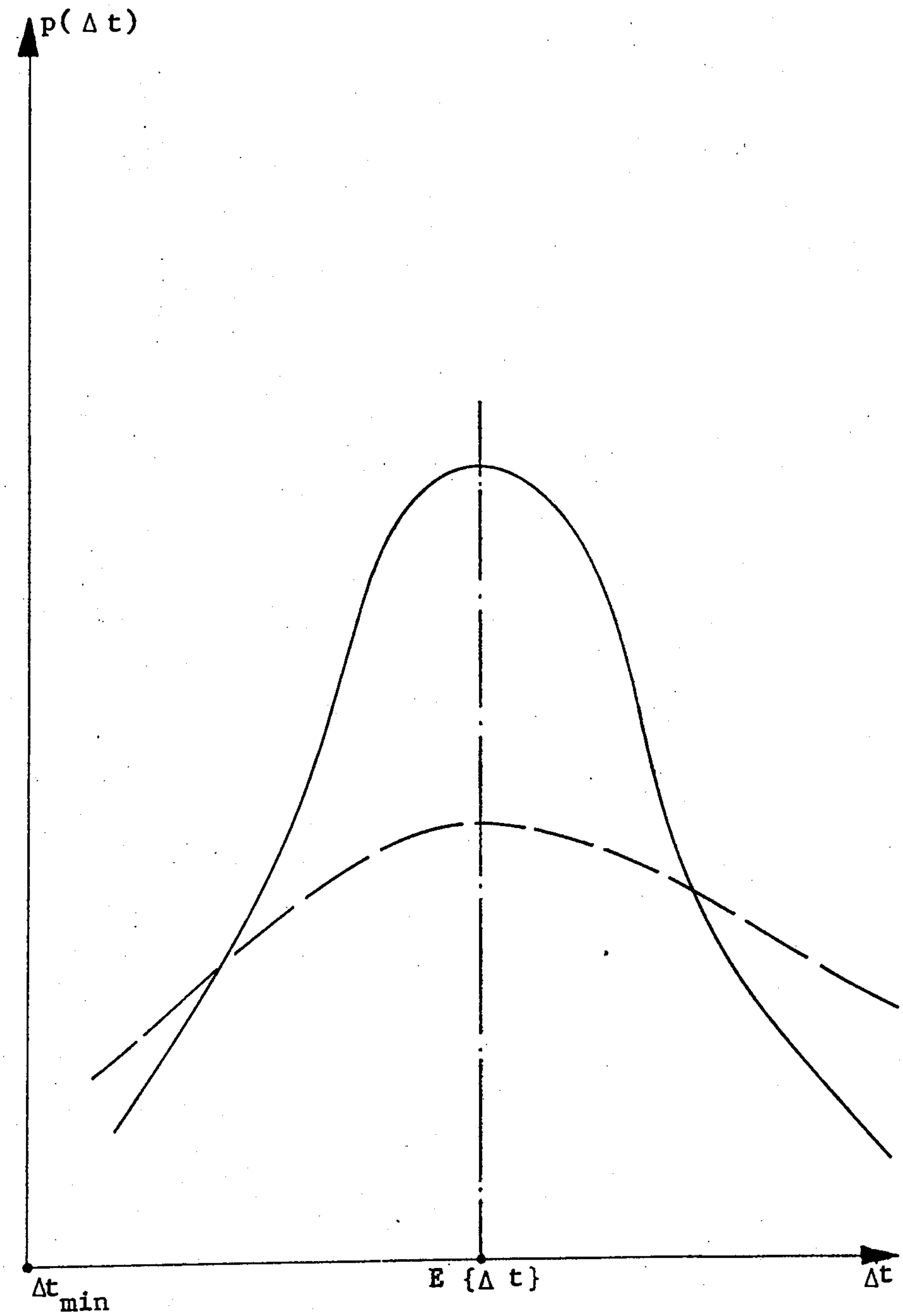


Fig. 7.1 POSSIBLE SHAPES OF p(Δt) FOR VARIOUS SOURCE SIGNALS.

the transmission frequency will be chosen $\frac{\Delta t_{j\,max} - \Delta t_{j\,min}}{2}$ for the optimum result. Obviously signals having these characteristics will not be suitable for the VTSIP system. But with these type of signals entropy will be maximum. From this discussion it is evident that maximum entropy will not mean any profit for our system. As the entropy of the system increases the store needed will be larger, so will be the system delay. Therefore, signals having small entropies will be easier to use with our system. Of course, when deciding whether to apply VTSIP or not, besides the entropy, the merit factor which is mentioned in chapter 6 must be considered.

Band limited random signals with stationary characteristics, like speech signals, are suitable for our system. Besides the speech signal, television, picture transmission, and even electrocardiograph signals can be transmitted using VTSIP by obtaining considerable bandwith compression, without sacrificing quality. The only cost paid is the system delay because of the store which is employed. In particular one-way communication systems can utilise high values of bandwidth reduction, as the time delay is not then important.

After the examination of the statistical characteristics of the signal, the standardization of the signal amplitude is necessary. The same problem also appears in the PCM systems.[5] Therefore, at this point, we can take advantage from the experience of the investigators who worked on these problems.

CHAPTER 8

EXPERIMENTAL RESULTS FOR SPEECH AS THE SOURCE SIGNAL

CHAPTER 8

8.1 EXPERIMENTAL RESULTS FOR SPEECH AS THE SOURCE SIGNAL

So far some theoretical investigations have been carried out. At the same time, while doing research on the system in general, an experiment has been designed to explore the practicalities of the system for a particular input signal. Before such an experiment can be designed, certain properties of the input signal must be known. For the speech signal chosen, no published information of the kind required was available. It was therefore necessary, before further practical realization of the system could be contemplated, to investigate these properties experimentally. Some attention was paid to the design of these experiments, as the results determine the required store size, and will thus considerably influence the cost of the system.

To produce the source signal, a tape containing several high quality telephone conversation recordings (supplied by the Post Office Research Station),was used throughout the experiments. Because of the high cost of the store, we have designed a simulation system rather than building a full scale telecommunication system. Details of the circuits used are given in the next sections of this chapter.

In the system each level change of the signal is detected, and a pulse is generated every time these changes occur. Each pulse will be used to initiate a word carrying information about the time interval between this level change and the previous one. To investigate the time of occurrence of these level changes, a reversible counter was used to represent the store. Every time a level change is detected, a pulse is applied to the 'UP' count input. At the same time an oscillator, with frequency equal to the transmitted pulse group frequency is connected to the 'DOWN' count input. Therefore, the reversible counter indicates the instantaneous occupancy of the store, which is continuously recorded on a paper tape, enabling us to register the store occupancy change with time.

We had two separate parameters which we were able to vary as we like, namely the transmission rate and the store size. By varying these two parameters their effects on the overflow and underflow has been investigated. A number of values of store size were now chosen, representing the range over which the store was likely to be usable. A second counter was then arranged to count the number of level changes occurring during the times when the count number in the reversible counter exceeded the store size set. This count therefore is a measure of those input level changes which would be absent in the reconstituted signal, and thus represents 'overflow distortion'. By counting these pulses and dividing their total number by the total number of pulses produced by the level change detector, a proportion is obtained. We have called this proportion the percentage overflow.

Underflow distortion was indicated by a third counter, connected to count the number of transmitted pulse groups occurring during the times when the store was empty. This number, when divided by the number of transmitted pulse groups will give the percentage underflow. At first sight this might seem to be a time percentage, but this is only because the transmitter oscillator is generating a constant frequency. Actually it is the amount of information which is redundant.

In the experiments linear quantizing is employed. Therefore, by reducing the input amplitude, we had the opportunity to make the same measurements for a reduced number of amplitude quanta.

As the rate of information at the input to the store is dependent on the amplitude of the signal, then at the beginning of every experiment the amplitude had to be standardized. As mentioned in the previous chapters, the amplitude of the speech signal is adjusted to be 14 dB below the peak clipping level. To obtain this accurately the digital output of the level change detector has been observed. When all digits were 1 or 0, a peak clipping occurred. Therefore, when these observations were less than 1% of the time this level is accepted as the normal signal level which is 14 dB below the peak clipping level. At the beginning of the prerecorded tape, a signal of 1 kHz frequency with almost constant amplitude has been recorded. The amplitude level of the 1 kHz signal which corresponds to the above mentioned speech amplitude level has been measured. For every experiment, level adjustments have been made referring to this 1 kHz signal.

With the first series of experiments, the results have been obtained for both male and female voices as the source signal. By using these results the curves in Fig.8.1 and Fig. 8.2 were obtained. This first series of experiments helped us to calibrate our system for a subsequent and more accurate series of experiments.

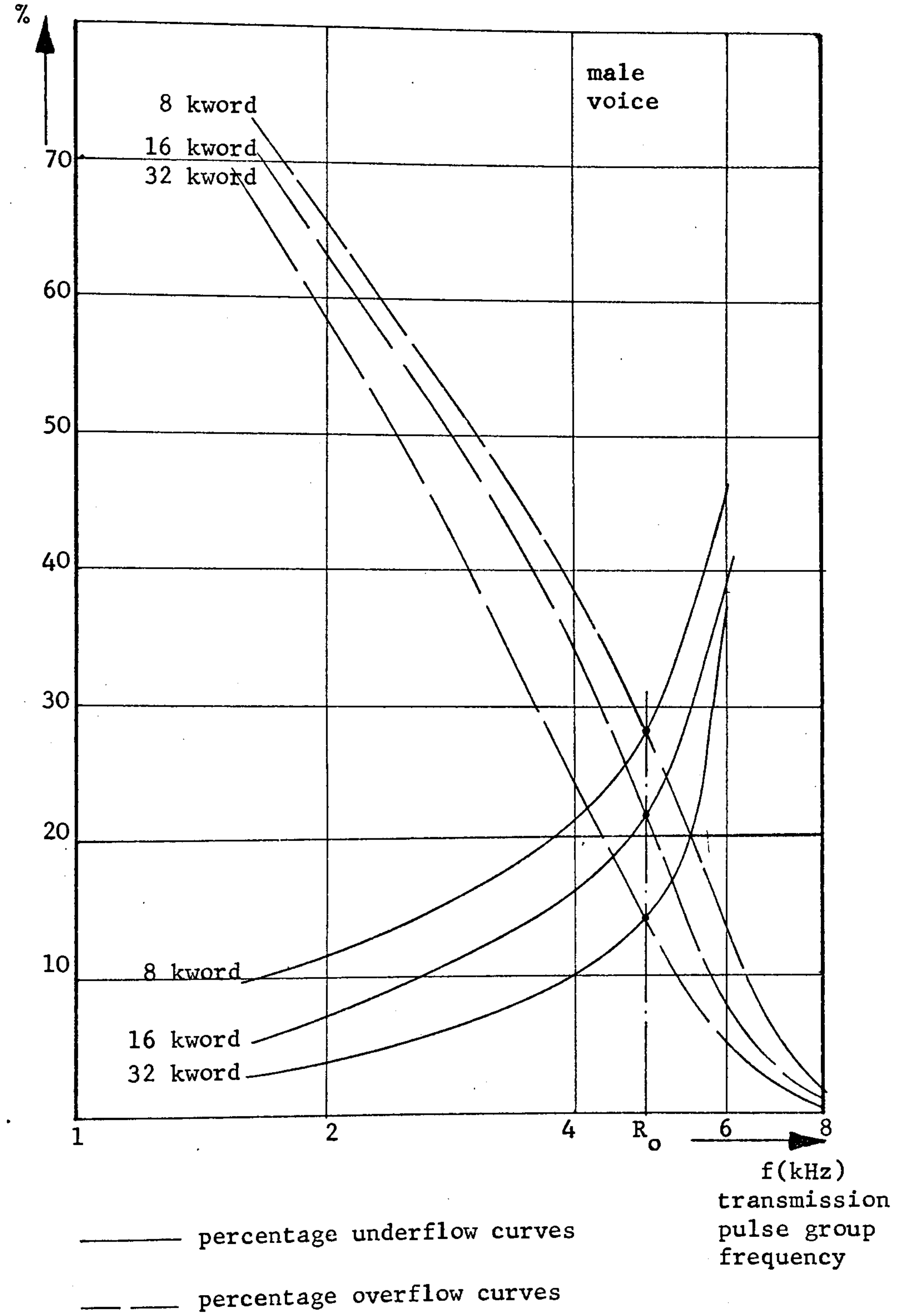


Fig. 8.1 PERCENTAGE OVERFLOW AND UNDERFLOW FOR VARIOUS TRANSMISSION FREQUENCIES AND STORE SIZES.

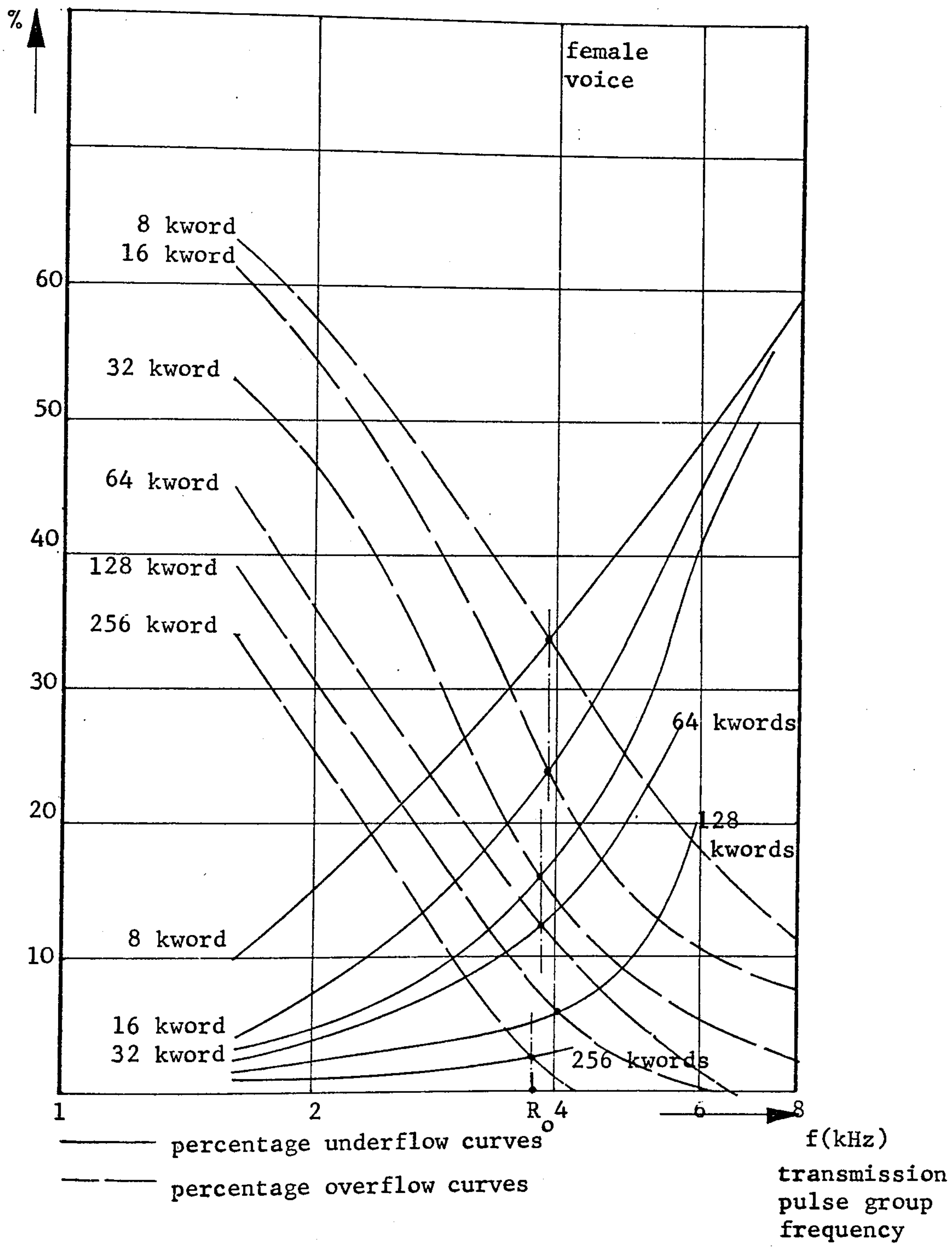


Fig.8.2 PERCENTAGE OVERFLOW AND UNDERFLOW FOR VARIOUS TRANSMISSION FREQUENCIES AND STORE SIZES.

With this second series of experiments, it became clear that there was little difference between the results using male and female voices, provided that they were accurately standardized.

The types of curves drawn are, percentage overflow and underflow against transmission frequency for various store sizes. Besides these measurements, the total period of overflow has been measured. If these values are divided to the total speech period, percentage overflow time can be found. In Figs. 8.3a, b, and c, these results have been shown.

From the curves given in Fig.8.3, we can see that as the frequency increases, the percentage overflow decreases, but the percentage underflow increases for a particular store size, as expected. An interesting point is that both curves cross each other at almost exactly the same frequency for different store sizes. As the transmission frequency has been taken to be the frequency at which the percentage overflow and underflow are equal, this result comes as no surprise, as this independence from the store size can be seen from the expression in equation 5.23.

A further set of curves is given in Fig.8.3a, b, and c, these are the graphs showing the percentage of time that the overflow condition exists. It will be noted that these curves lie well below the corresponding curves representing percentage overflow of the store. This would indicate that the overflow condition exists mainly during these times when the signal voltage is changing most rapidly. Thus, although for the smaller store sizes the percentage of lost information caused by the overflow mechanism is high, this represents only a relatively low percentage of time during which the information is being lost.

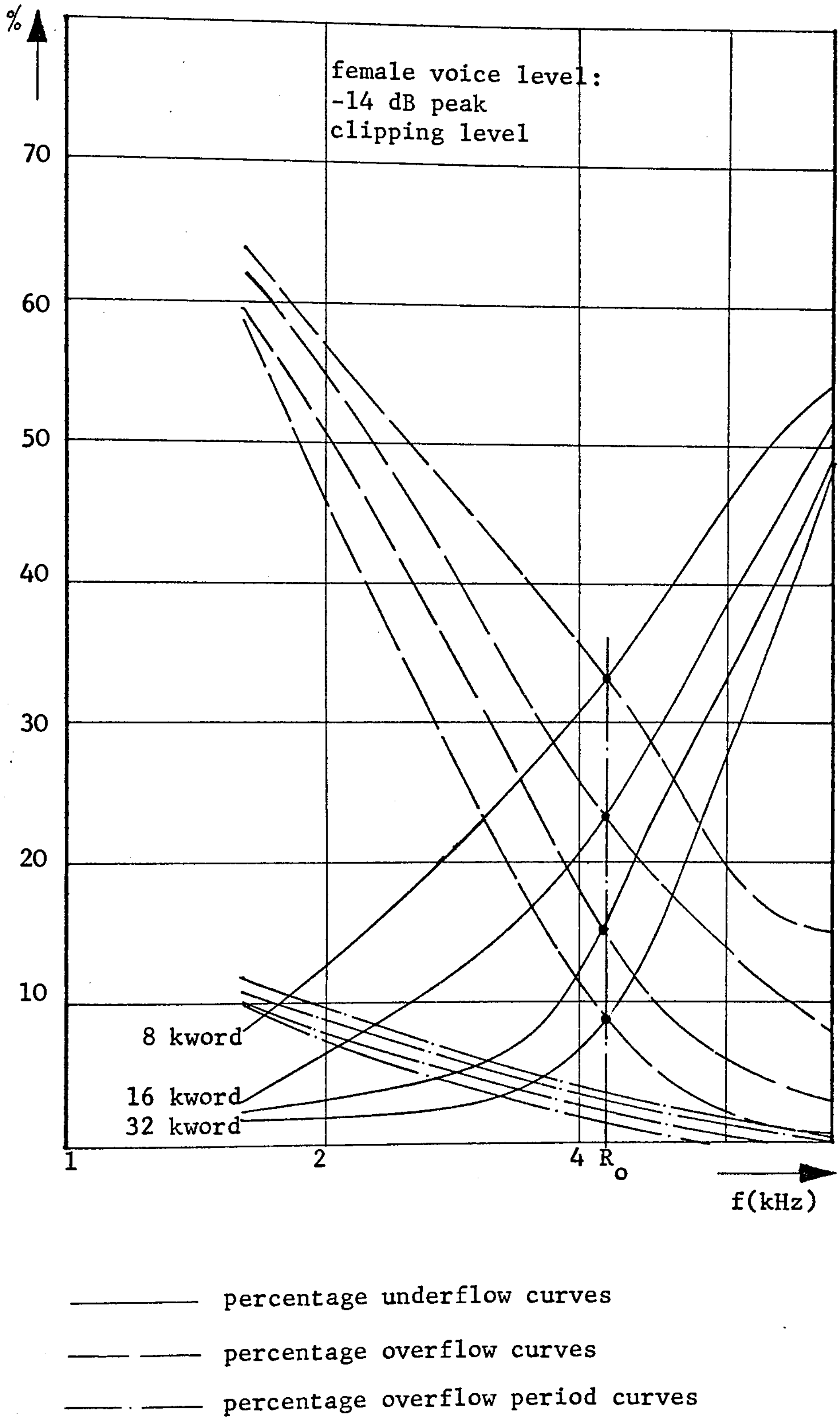
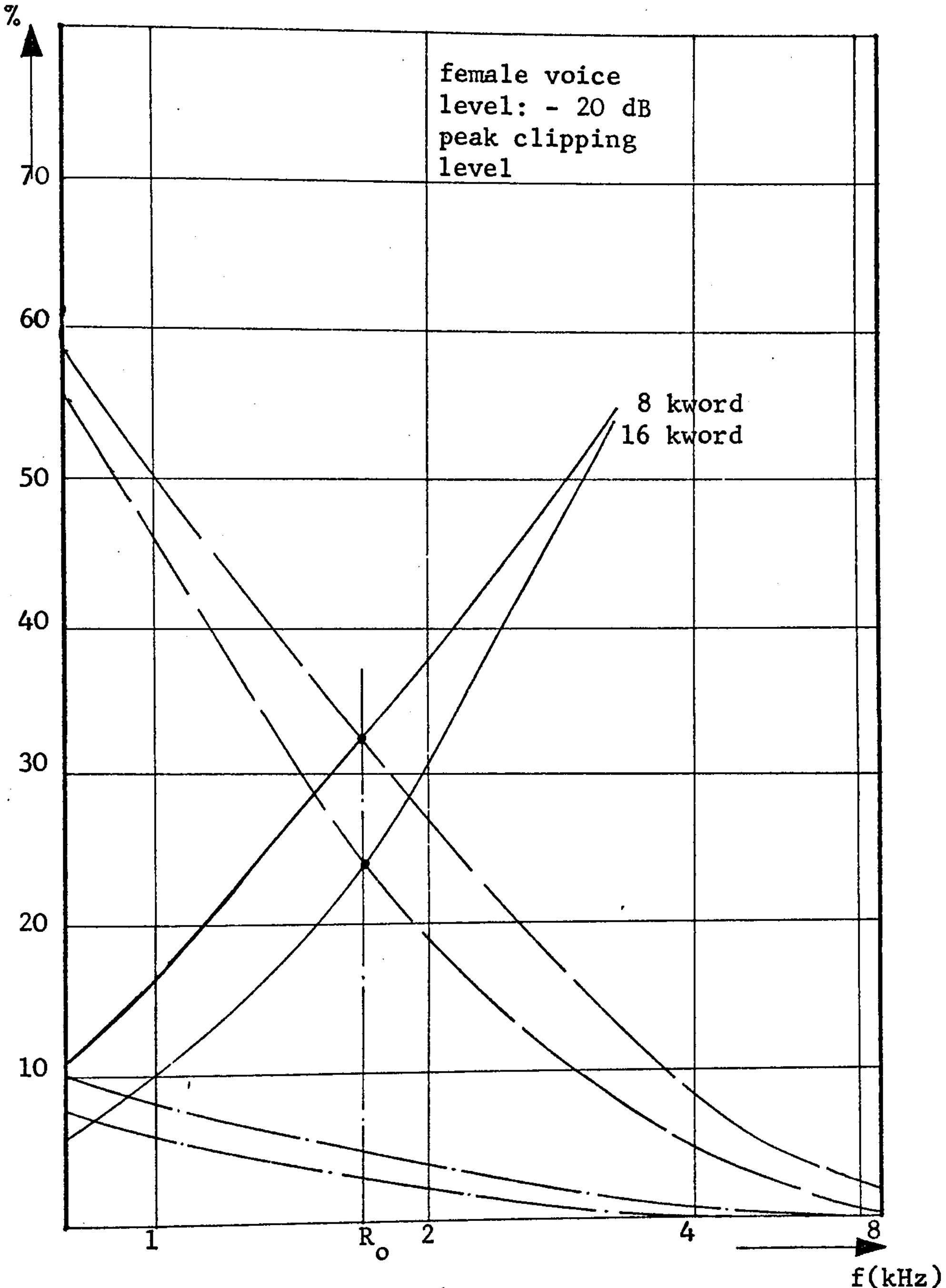


Fig.8.3a. PERCENTAGE OVERFLOW, UNDERFLOW AND OVERFLOW PERIOD CURVES FOR VARIOUS STORE SIZES AND TRANSMISSION FREQUENCIES.



——— percentage underflow curves

— — percentage overflow curves

—·— percentage overflow period curves

Fig. 8.3b PERCENTAGE OVERFLOW, UNDERFLOW AND OVERFLOW PERIOD CURVES FOR VARIOUS STORE SIZES AND TRANSMISSION FREQUENCIES.

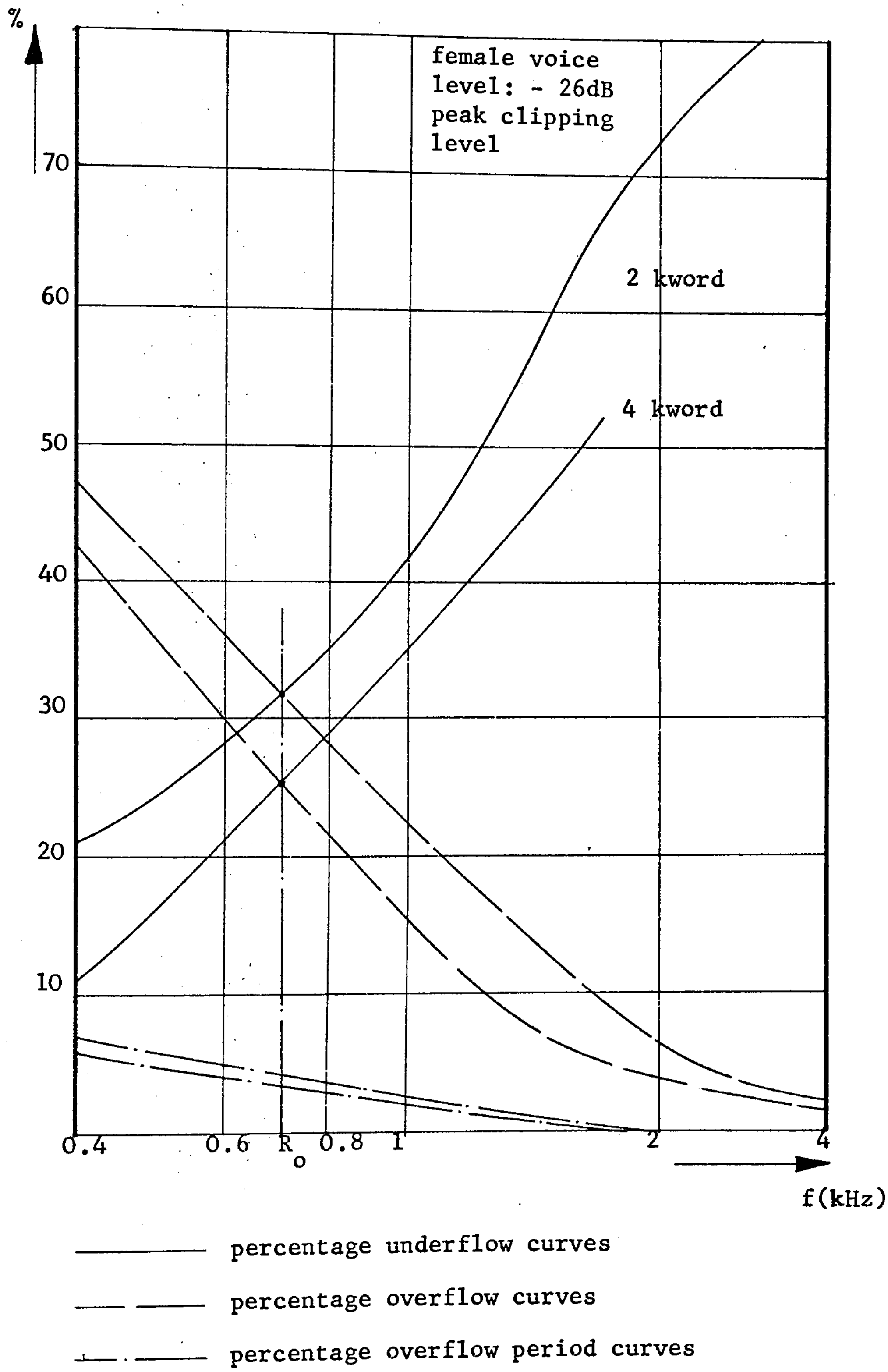


Fig.8.3c PERCENTAGE OVERFLOW, UNDERFLOW AND OVERFLOW PERIOD CURVES FOR VARIOUS STORE SIZES AND TRANSMISSION FREQUENCIES.

A full investigation of the subjective effect of this loss on a speech signal will have to be undertaken sometime in the future. However listening tests using the tapes on which had been recorded the distorted speech signals, gave the impression that a high level of percentage overflow could be tolerated. Overflow distortion appeared as 'clicks' added to the signal. When the overflow rate was high, the noise which was introduced by the clicks increased. But, because of the high tolerance of the human ear to the sounds to which it is familiar, the intelligibility of the speech was not seriously reduced. We did not attempt to carry out subjective quality tests, as long periods of conversation have been recorded which would not enable a human being to make an objective assessment about the average quality of the individual conversations. With the experiments which have been carried out so far, we have only attempted to verify the practicalities of the proposed system. The details of the measurement procedure is given in section 8.2.

Another way of finding optimum frequency is to count the total number of samples and take the average over the speech period. When this was done, almost equal values to that found from the curves were obtained. For example, for the first experiment in the second set, the total number of samples was approximately 1.300.000 with a speech period of approximately 315 secs. and the optimum transmission frequency $R_o = \frac{1.300.000}{315} \simeq 4.150$. This compares well with the 4.200 obtained from the curves. This result is not surprising, as the total number of samples - lost samples because of overflow = total number of transmitted pulse groups - redundant pulse groups.

In an ideal system, the number of lost samples and the number of redundant samples would be zero. In this case the total number of samples would be equal to the total number of transmitted pulse groups. At the optimum transmission frequency these two will be equal again as this time the number of lost samples will be equal to the number of redundant pulse groups. This relation is verified by the experiments.

In addition to the measurement of overflow and underflow, a continuous recording was made of the store occupancy over the period of the experiment. This enables an inspection of the storage process to be made. A sample of this recording is shown in Fig.8.4.

The system introduces a delay because of the store used. This delay can be calculated by using the results of these experiments. A curve showing time delay against store size for the chosen optimum transmission frequency is plotted in Fig.8.5. If the optimum frequency is R_o and the store size is S_{max} the delay Δ_{VTSIP} is

$$\Delta_{VTSIP} = \frac{S_{max}}{R_o} \text{ secs} \qquad 8.1$$

The maximum tolerable delay can be found after making subjective tests. This was not attempted in this research project.

8.2.1 Experimental Methods

To carry out the experiments which have been mentioned in Section 8.1, a measurement system was developed. To keep the expense to a minimum we used available instruments wherever possible. We had to count pulses coming from five different sources to be able to derive the necessary information. These were :

(1) The total number of samples (n_s)

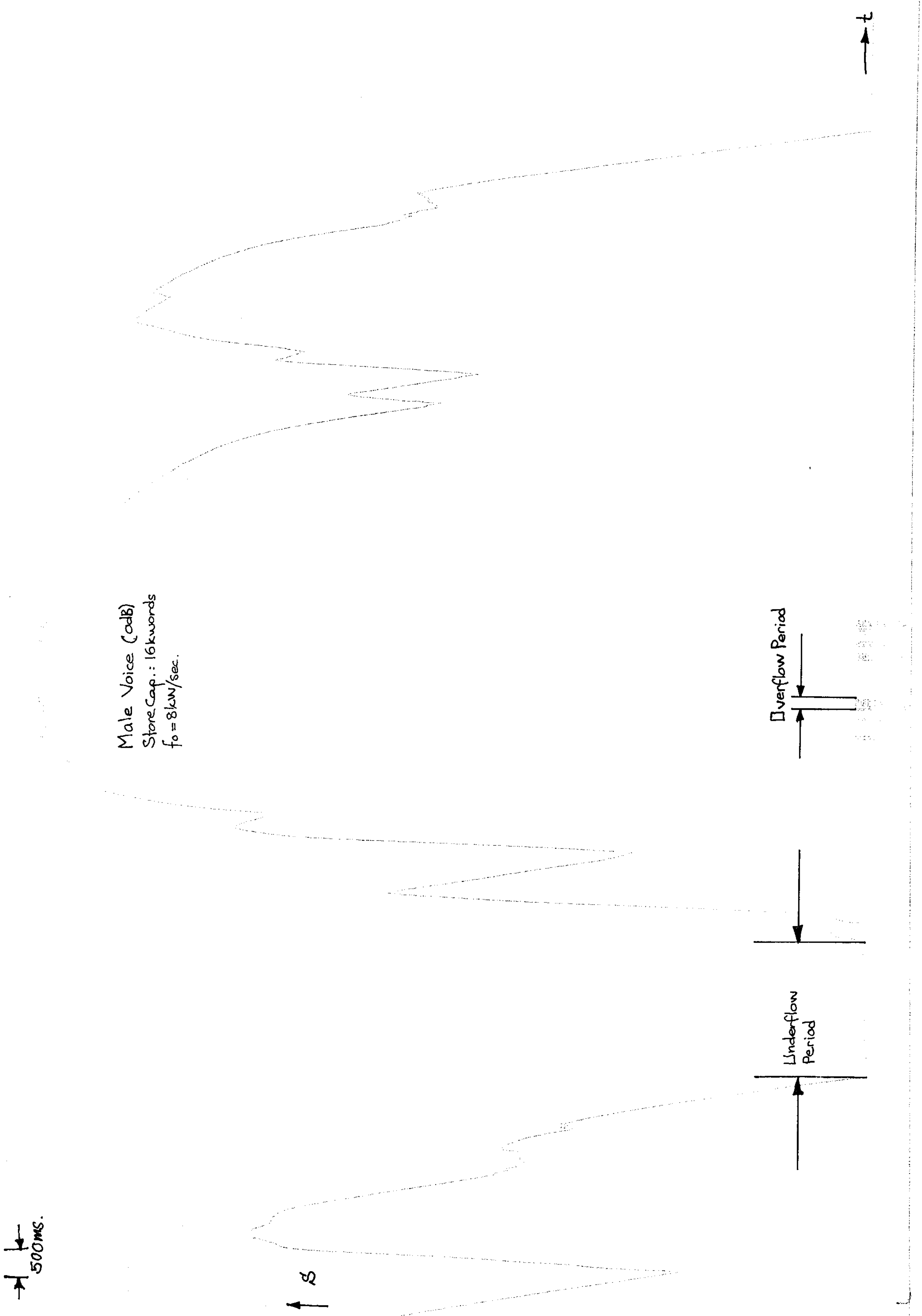

Male Voice (cdB)
Store Cap.: 16kwords
f_0 = 8kw/sec.
Overflow Period
Underflow Period
500ms.
S
t

fig.8.4.

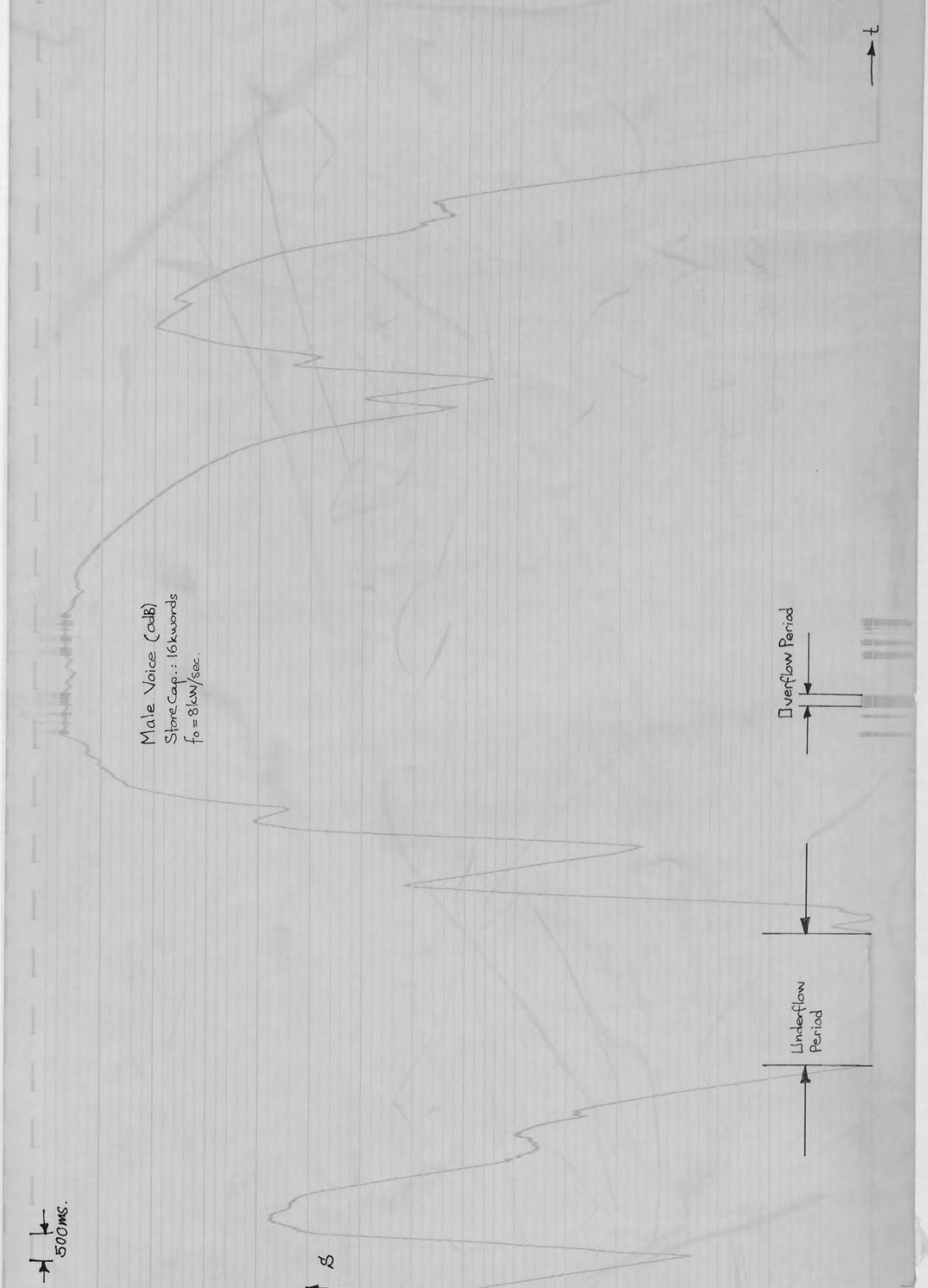

500ms.
Male Voice (odB)
Store Cap.: 16kwords
f_0 = 8kW/sec.
Overflow Period
Underflow Period
s
t

fig.8.4.

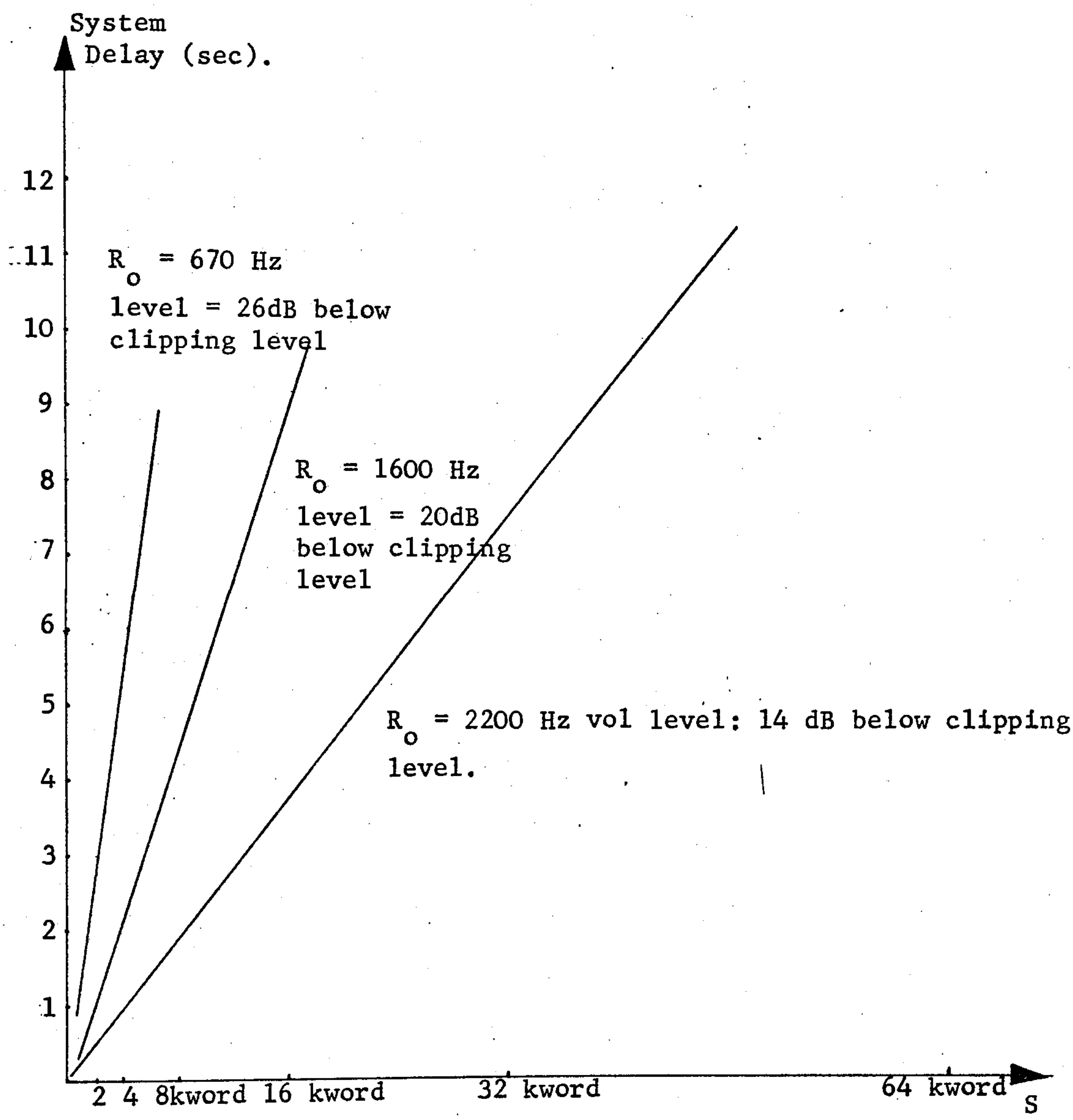


Fig. 8.5 SYSTEM DELAY

(2) The total number of lost samples, because of the overflow (n_{of})

(3) The total number of transmitted pulse groups (n_t)

(4) The total number of transmitted useless pulse groups (n_r)

(5) The total length of the overflow period (T_{of}).

These were measured for a speech period of approximately five minutes. The measurement system used is shown in Fig.8.6. In addition to those measured, the overflow distortion was simulated on the actual speech signal and recorded on a tape. In the meantime the store occupancy was continuously recorded on a paper tape using a high-speed recorder.

A device, 'the level change detector', has been developed to detect the instants when the signal crosses the predetermined amplitude levels. The detailed information is given in section 8.2.2. This device givesa 50 ns pulse everytime the signal crosses a predetermined amplitude level. There are two outputs from this device, one gives pulses of increasing and the other for decreasing signal amplitudes. An 'OR' gate combines these pulses. Every pulse is assumed to represent a word carrying information about the length of time interval between the previous pulse and itself. Therefore, these pulses occur with the average period of $\Delta t_{j\ av}$. Each of these pulses adds 1 word of information to the store. A reversible counter is used to represent the store. This reversible counter has been designed and constructed in the University of Aston, Department of Electrical Engineering, as an M.Sc. project[12]. The reversible counter is able to count up to a maximum frequency of 500 kHz regardless of pulse shape and time of occurrence. 'Up' and

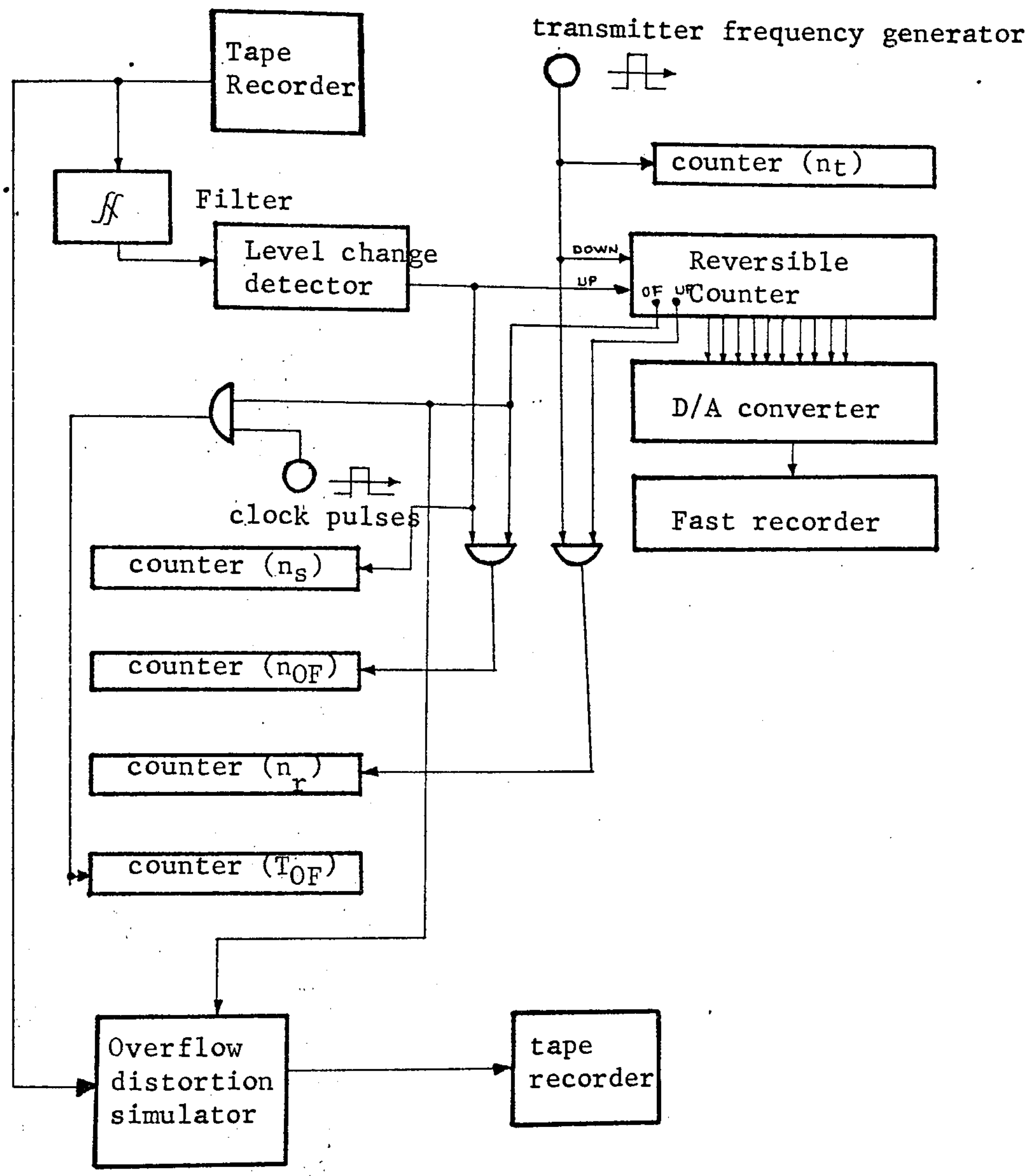


Fig. 8.6 MEASUREMENT SYSTEM.

'Down' count inputs can accept pulses of any duration and both can count at the same time. The pulses coming from the output of the OR gate are applied to the 'Up' input to the reversible counter. A stable oscillator with adjustable repetition frequency is used to generate the pulses representing the transmission frequency. These pulses are applied to the down input of the reversible counter. The reversible counter continuously shows the occupancy of the store capacity. The reversible counter is 16 binary digits. Therefore, it is possible to arrange the store size between one word to approximately 131000 words. By means of logic gates the overflow and underflow conditions have been detected as shown in Fig.8.6. When an overflow has occurred the up count input to the reversible counter is blocked, prohibiting information to be registered in the reversible counter. At the same time a gate has been opened allowing these pulses to be conveyed to the input of 'the total number of lost samples' counter. This condition persists until the next transmitted pulse group is generated, when the system returns to its normal state. At this instant it is possible to see the number of lost samples from the indicators of the relevant counter.

When the underflow condition occurs this is sensed by the logic circuits. This time another gate permits the transmission frequency oscillator pulses to reach the input of the 'useless number of transmitted pulse groups' counter. On the termination of the underflow period this gate again blocks the way of the input to this counter. Another counter is connected to the output of the transmission frequency oscillator so that it indicates the total number of transmitted pulse groups.

Besides counting the total number of lost samples during the presence of overflow a crystal controlled 1 MHz pulse generator output is connected to the input of a further counter to count the total overflow period in multiples of μs.

As explained in section 4.3.2 an overflow distortion occurs in the reconstituted signal because of the store size limitation. The graphical representation of this distortion is also shown in the same section in Fig.4.3. With this series of experiments we planned to test the subjective effects of this overflow distortion. To do this, the speech signal is applied to a differentiator circuit followed by an integrator circuit, causing no change in the signal during normal working conditions. When the overflow occurs a transistor is activated and disconnects the integrator from the differentiator output as shown in Fig.8. 7. The charged integrator capacitor retains the charge and avoids any level changes at the output of the integrator, until the overflow disappears. When the overflow is terminated, the transistor reconnects the output of the differentiator to the input of the integrator. Since the output voltage of the integrator cannot make a sudden jump to a new value, it follows the input signal with a certain d.c. level shift. This shift is only temporary, however, as the resistor connected in parallel with the integrator's capacitor ensures that the mean voltage across the latter, taken over a long period, is zero. The output signal from the integrator is applied to the input of the tape recorder.

The output of the reversible counter is connected to a 10 bit digital to analogue converter. This D/A converter is also another

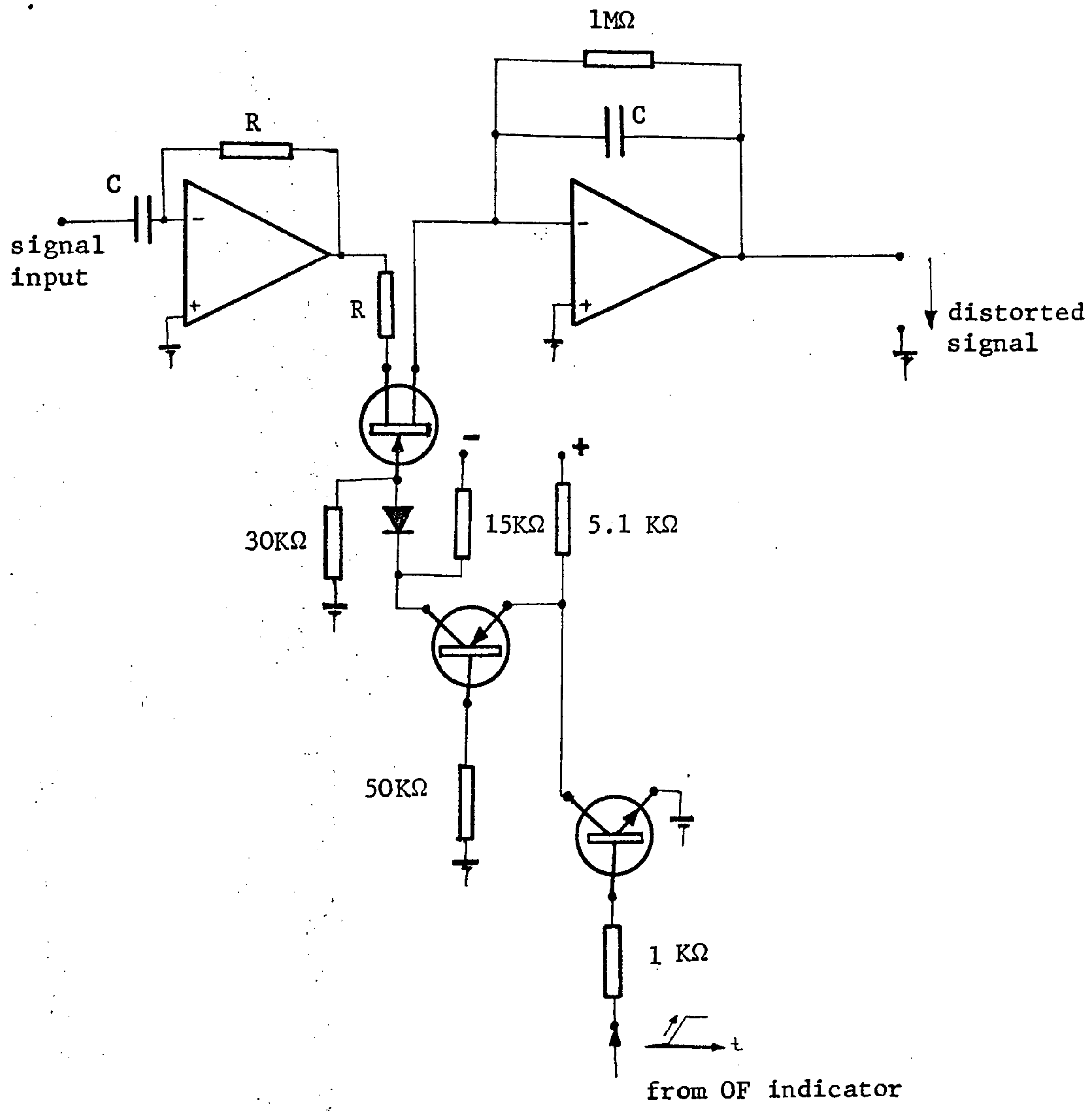


Fig. 8.7 OVERFLOW DISTORTION SIMULATOR.

M.Sc. project[16]. The analogue output of the D/A converter is connected to the high speed recorder, which is arranged to have a full scale deflection corresponding to the maximum capacity of the store used.

The input signal is obtained from the tape as mentioned in Section 8.1. To limit the signal bandwidth a filter is employed between the tape recorder and the input of the level change detector.

As the reversible counter cannot count the frequencies above 500 kHz, then instead of doing the experiment in real time we reduce the tape speed to one half of the actual recording speed.

The curves in Fig.8.1 have been prepared by calculating percentage overflow from n_{of}/n_s, and percentage underflow from n_r/n_x. Percentage overflow period has been calculated from $T_{of}/n_x.R_o$.

8.2.2 Design of the Amplitude Change Detector

This device has been designed to enable us to detect the instants when the source signal crosses the predetermined amplitude quanta levels. The device has been planned to have 64 uniformly distributed amplitude levels on each side of the zero axis. To observe the digital quantized value of the signal 7 digital outputs are available. Outputs giving increasing or decreasing level changes are provided.

To be able to detect the level changes in both directions two comparators are employed. To avoid any instabilities because of the high gain of these comparators and to increase their noise immunities, the unused inputs have been connected to the voltages which are equal to the half quantum value. Synchronous working has been found most suitable for our purpose.

The design is basically a feedback system. A change of more than one quantum in the input voltage in either direction causes a level change in the output of the relevant comparator, causing an 'AND' gate to open, letting in the clock pulses. The first clock pulse is applied to the appropriate input of a binary reversible 8 bit counter, causing 1 bit change at the digital output of the counter, according to the direction of the signal. By using a D/A converter a current proportional to the amplitude of the binary output of the counter is obtained. This current is then fed back to the same point as the input current, so that it neutralises the change of this current.

To avoid unwanted feedback, by using several monostable multivibrators, the channel which is not active is blocked until the system reaches its steady state. This period has been calculated as :

$$\text{device settling time} = T_{pd_{comp.}} + T_{pd_{gates}} + T_{pd_{counter}} + T_{pd_{D/A con.}} + T_{D/A\ settle}$$

where T_{pd} propagation delay of the device.

The periods of the monostable multivibrators are arranged to be longer than this settling time. The comparator outputs proved to be noisy, so a Schmitt-Trigger circuit was employed as the logic gate, as seen in Fig. 8.8.

When all the levels have been crossed, then as the D/A converter cannot supply any more current, one of the comparators cannot change its state, and further clock pulses will not pass through the circuits. To avoid this occurrence, a logic circuit has been designed to detect when all binary outputs of the reversible counter are either 1 or 0. When this condition exists the clock pulse path is blocked.

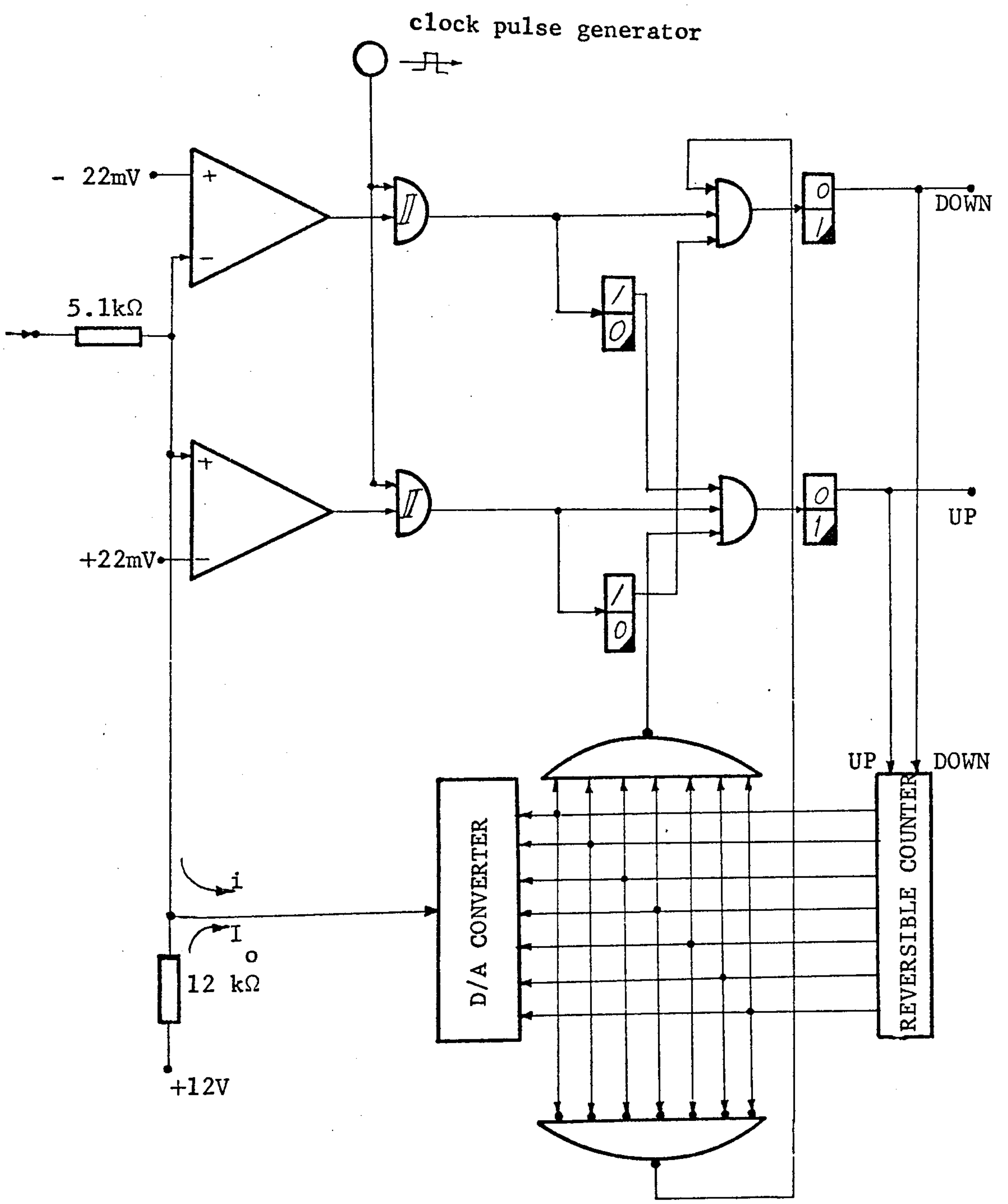


Fig. 8.8 LEVEL CHANGE DETECTOR

When the input voltage is zero, the D/A converter output current is - 1mA. To compensate for this current, a resistor is connected to the positive voltage supply, so bipolar operation is thereby obtained.

The outputs, giving the pulses when the signal level changes, are obtained from the corresponding inputs to the reversible counter.

CHAPTER 9

CALCULATION OF THE SYSTEM PARAMETERS USING SPEECH AS THE INPUT SIGNAL

CHAPTER 9

9.1 CALCULATION OF THE SYSTEM PARAMETERS USING SPEECH AS THE INPUT SIGNAL

When speech is used as the input signal to the VTSIP system, we can define the system parameters for this particular case. With the experiments the variations of these parameters have been investigated.

If we accept that speech signals have a frequency band of 300 Hz - 3400 Hz, then referring to section 3.3.3, we can calculate the necessary word length for the time interval coding. From equation 3.5 the code length is:

$$l_c = \log_2(\Delta t_{max} / \Delta t_{min}) \simeq 11 \text{ bits/word} \qquad 9.1$$

Therefore, a word consisting of 11 bits should be used for the time interval coding. Knowing this, if we can calculate the optimum transmission frequency R_o, then the channel capacity R can be calculated.

Using equation 6.8 :

$$R = \frac{l_c}{\Delta t_{av}} = \frac{11}{\Delta t_{av}} \text{ bits/sec} \qquad 9.2$$

where $\Delta t_{av} = \frac{1}{R_o}$

The value of Δt_{av} can be found from equation 5.23. Before doing this calculation we must find the first derivative density function of the speech signal. Some research in determining this has already been done[29]. In this work, the amplitude of a speech signal was sampled at regular intervals of 125 μs. Taken over a 5 minute period, it was found that for 67% of the time the amplitudes of successive samples were equal, i.e. for only 33% of the time was there a change between successive samples.

We therefore write,

$$T_{active} = T_{speech} \times 0.33$$

where,

T_{active} = time during which a change occurs

T_{speech} = period over which measurements have been made to find the characteristics of the speech signal.

In these experiments, T_{speech} was 300 sec. Thus T_{active} = 99 secs.

A further result of this research was that the expected value of amplitude differences between successive samples is 1.65 quantum.

From these results we can find the total number of level changes for a speech signal in the speech period:

$$n_s = \frac{(T_{speech} \times 0.33)}{125 \ \mu s} \times (1.65) = 1.31 \times 10^6 \text{ samples} \qquad 9.3$$

This value of course, is for a speech amplitude level which is adjusted to be 14 dB below the peak clipping level. To find the average time interval:

$$\Delta t_{av} = \frac{T_{speech}}{n_s} \simeq 228 \ \mu s/\text{sample} \qquad 9.4$$

From t_{av} optimum transmission frequency :

$$R_o = \frac{1}{\Delta t_{av}} \simeq 4{,}380 \text{ Hz} \qquad 9.5$$

As we know, in Pulse Code Modulation, 8 kHz is the transmission pulse group frequency, and each pulse group consists of 7 bits, therefore:

$$C = 7 \times 8000 = 56 \text{ kbits/sec.}$$

Then the merit factor: $M = \frac{C}{R_o} = \frac{56000}{4{,}380X11} \simeq 1.17$ 9.6

This result shows that by using the VTSIP system the same signal can be transmitted over a channel having less capacity. Of course, when reconstituted, the signal which is transmitted by VTSIP will be less affected by the quantizing noise because of the reasons explained in section 5.1. Referring to this section, we know that instead of using 128 amplitude levels, we can obtain the same signal to noise ratio by using 32 levels. In this case obviously Δt_{av} will be:

$$\Delta t_{av} = 228 \times 4 = 912 \ \mu s$$

$$\text{and} \quad R_o \simeq 1095 \text{ Hz}, \ 1_c = 9$$

assuming that linear quantizing is made. This will give a merit factor of:

$$M = 5.18 \qquad 9.7$$

which represents a drastic reduction in the channel capacity which is occupied by the VTSIP signal continuously. Obviously this result needs experimental verification which it has not been possible to carry out in the time available for this research work.

We will now calculate the distortion caused by the finite size of the store. If overflow exists, it will continue for Δt_{av} seconds. Therefore, we have to know the average number of level changes which exist in this period. The average level change in 125 μ_s is 1.65 quantum. Therefore, in Δt_{av} seconds:

$$d_{OF} = E\left[N_x\right] = \frac{\Delta t_{av}}{125\mu s} \times 1.65 \simeq 3.02 \text{ level changes} \qquad 9.8$$

As there are 128 levels altogether, then if the percentage distortion has to be calculated :

$$D_{OF} = \frac{3.02}{128} = 2.35\%$$

As an average, approximately 3 level changes occur during each overflow period. In the experiments described in section 8.1

we have measured the total length of the overflow period, T_{OF}; this was 13.5 secs, when an 8 kword store size was used. Every occurrence of overflow caused a loss of information corresponding to 3 level changes. Therefore, as one overflow occupies Δt_{av} seconds, the number of lost samples will be :

$$n_{OF} = \frac{T_{OF}}{\Delta t_{av}} \times N_{x_{av}} = 178{,}500 \text{ samples} \qquad 9.9$$

The percentage of lost samples as defined in section 4.3.2 is :

$$\text{percentage overflow} = \frac{n_{OF}}{n_s} \simeq 13.6\% \qquad 9.10$$

The acceptability of this figure has been discussed in section 8.2.1

Referring to the curves in Fig.8.3a, we can compare the results of these calculations with the results obtained from the experiments. For the 8 kword store size and speech amplitude at 14 dB below the peak clipping level $n_s = 1.31 \times 10^6$ samples have been counted. This was equal to that found by using the data taken from Whitehouse's work[29]. From the curves the optimum transmission frequency is found to be 4150 Hz instead of 4380 Hz, which is found from the calculations. This also shows that the calculations compare well with the experimental results.

If the transmission frequency is chosen to be equal to that defined as the optimum, the percentage underflow will be equal to the percentage overflow.

When the system parameters have been found the system delay is then:

$$\Delta_{VTSIP} = \frac{S_{max}}{R_o} = \frac{8 \text{ kword}}{4.15} \simeq 1.93 \text{ secs.} \qquad 9.11$$

Besides these calculations, if we try to find the noise introduced by the time quantizing process using equation 5.20, we find that this noise component becomes negligibly small with respect to the amplitude quantizing noise. To find the irregularities in the quantizing process is a further step which must be carried out when designing the actual system.

CHAPTER 10

CONCLUSIONS

CHAPTER 10

10.1 CONCLUSION

For many years much research effort has been devoted to a reduction in the bandwidth needed for PCM transmission. The present research adopts a different approach to the problem. By the use of a buffer store, the transmission bandwidth required is that corresponding to the mean, rather than the maximum rate of information contained in the signal. The use of the store, however, necessarily results in a time delay of the signal in the process of transmission. In addition, because of the finite store size a form of distortion is introduced caused by a loss of information when the store is filled.

It is envisaged that the system developed may have applications involving speech signals, and a considerable part of this research has had to be concerned with an analysis of those properties of speech which are important to the design of the system, and about which no information previously existed. In particular, a detailed knowledge of certain properties of speech must be obtained before the required store size can be determined. This information could not be obtained until experiments had been conducted, the results of which were not obtained before a late stage in the research, and this, together with the high cost of the store, prevented the construction of a complete transmitter-receiver system. However, apparatus was made to simulate part of the system, enabling a subjective assessment of the distortion to be made.

A feature of the new system is the improvement on the ratio of signal to quantizing noise, when compared with the standard

PCM system. Much further work on this aspect of the system remains to be done. Investigation is needed into the optimization of the signal to noise ratio by variation of the number of quantized amplitude levels. Also, it is expected that a substantial improvement in signal to noise ratio could be obtained from an examination of the effects of non-linear quantization both of the amplitude levels and of the time interval measurements.

Another promising line of investigation is the possibility of reducing the store overflow distortion by arranging that the store occupancy affects the transmitted pulse frequency. It is likely that a small variation in this frequency would result in an appreciable reduction in the distortion.

An examination of the results of the speech analysis reveals the possibility that prediction of overflow is feasible. This should be investigated, as it may suggest a method of reducing the distortion caused by store overflow.

This research has been confined to a single-channel communication system. There is obviously no reason for suggesting that the principles could not be used in a multiplex system. In that case either a separate store for each channel could be used, or it might be possible to use a common store: there is clearly scope for the development of a method of economizing in the total storage capacity needed. A further line of investigation would be to consider the use of a common store for both transmitting and receiving functions.

Much work remains to be done, but the author feels that in three very enjoyable years spent in the University a new principle has been established which, when further developed, could well result in a significant improvement in communication systems.

APPENDIX = Simplification of the Integral equation to find p(S).

If we assume that $p_z(\frac{dy}{dt})$ is an exponential function as suggested by Whitehouse[29], to find the p(S), the integral equation in 4.14 can be treated as follows:

$$p(S) = \int_o^\infty \frac{aR_o}{(S_i+1-S)^2} \cdot p_z\left(\frac{aR_o}{S_i+1-S}\right) p(S_i)dS;$$

by substituting $p_z\left(\frac{aR_o}{S_i+1-S}\right) = A\, e^{-B\frac{aR_o}{(S_i+1-S)}}$ A.1

where A, B are constants,

$$p(S) = \int_o^\infty \frac{aR_o}{(S_i+1-S)^2} A.e^{-B\frac{aR_o}{(S_i+1-S)}} \cdot p(S_i)ds \qquad A.2$$

If we integrate A.2 by parts;[25,8]

$$p(S) = \frac{A}{B}\, e^{-B\frac{aR_o}{(S_i+1-S)}} p(S_i) \Big|_o^\infty - \frac{A}{B} \int_o^\infty e^{-B\frac{aR_o}{(S_i+1-S)}} p(S_i)dS_i \qquad A.3$$

By differentiating A.3 with respect to S, we obtain constant coefficient differential equation:

$$\frac{dp(S)}{dS} = \frac{A.aR_o}{(1-S)^2}\, p(o)\, e^{-B\frac{aR_o}{1-S}} + p(S) \qquad A.4$$

If this equation is solved p(S) can be found.

LIST OF SYMBOLS

$f(t)$	signal
$\delta(t)$	delta function
$G(t)$	signal after passing from the low pass filter
ω_s	sampling frequency
T_s	sampling period
$\bar{\omega}$	radial frequency
a	amplitude of unit quantum
m	number of quanta levels
I	information
f_1	lowest signal frequency
f_h	highest signal frequency
n	number of bits
Δt	time interval
A_j	jth amplitude level
Δ	delay
w	bandwidth of the signal
Δt_{min}	minimum time interval
Δt_{max}	maximum time interval
$y(t)$	signal
l_c	code length
k	constant
Δt_e	time interval measurement error
R_o	optimum transmission rate
S_i	previous state of the store
S	present state of the store
S_{max}	maximum store capacity
z_j	first derivative of jth amplitude level

N_x	number of levels in a certain period
d_{of}	overflow distortion
e_j	quantizing error for jth level
e	total quantizing error
VTSIP	Variable Time-Scale Information Processing
PCM	Pulse Code Modulation
$W(T)$	autocorrelation function.
$f(\omega)$	power spectrum
ω_Q	cut-off frequency of $f(\omega)$
τ	unit time quantum
Δa	amplitude error
$E\{\ \}$	expected value
$p(\ )$	probability density fn.
$\underline{P}(\ )$	probability fn.
Δv_j	encoder level shifts
Δy_j	decoder level shifts
$H(\ )$	entropy
R	information rate
M	merit factor
Δ_{VTSIP}	VTSIP system delay
n_s	number of samples
n_{of}	number of lost samples
n_t	number of transmitted pulse groups
n_r	number of redundant pulse groups
T_{of}	overflow period
T_{pd}	propagation delay
$T_{D/A}$	D/A converter settling time
T_{speech}	speech period
C	PCM channel capacity
D_{of}	percentage of distortion

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