SUITABILITY OF JUNCTION CABLES FOR PCM TRANSMISSION WITH MEASUREMENTS ON SELECTED KP & T CABLES.

THIS THESIS HAS BEEN ACCEPTED FOR
THE DEGREE OF M.Sc. 1979
AND A COPY MAY BE PLACED IN THE
UNIVERSITY LIBRARY

Mathias Kalanda Kizza, B.Sc. (Eng.)

A thesis submitted inpart fullfilment for the Degree of Master of Science in Engineering at the University of Nairobi.

UNIVERSITY OF NAIROBI LIBRARY
0145011 3

1979.

This thesis is my original work and has not been presented for a degree in any other University.

Mathias K. Kizza.

This thesis has been submitted for examination with my approval as University Supervisor.

Dr. A.V. Otieno.

ACKNOWLEDGEMENTS

The author wishes to acknowledge the many helpful discussions with and guidance from the supervisor, Prof. A.V. Otieno of the University of Nairobi's, Department of Electrical Engineering. The help offered by the employees of the K.P. & T. Corporation during the measurements presented in this report, was indispensable.

The author also wishes to acknowledge the scholarship offered by the Norwegian Government (through the Norwegian Agency for International Development), which made this work possible.

Many thanks also due to Mrs. Phyllis M. Njihia for rapid and accurate typing of the manuscript.

CONTENTS

ABSTRACT

An investigation into the suitability of paired, voice-frequency cables for PCM transmission and the relevant constraints involved, are given in this report. An outline of the basic concepts of a PCM system is given in Chapter One followed by a survey of the principal sources of interference in the transmission of digital signals over paired cables. It is found that near-end and far-end crosstalk and office noise are the only forms of interference that require specific engineering measures to combat. Supplementary factors such as line coding, cable characteristics and construction and operation of regenerative repeaters, that affect the development of line engineering methods are also discussed.

A theory is developed for predicting the interfering power from near-end and far-end crosstalk if a given number of 32-channel PCM systems operate in a cable. The properties of office noise are also studied. Using these facts, engineering procedures that are employed in determining the maximum number of PCM systems in any given cable so as to satisfy an error rate objective of 10^{-7} for a repeater section, are developed. Measurements of cable attenuation and crosstalk interference are then carried out on some selected Kenya Posts and Telecommunications

Corporation junction cables and their PCM cable-fill determined for various modes of operation. An indication on the suitability of each cable tested and a general cable evaluation procedure for future use are given.

SYMBOLS AND ABBREVIATIONS

p.a.m.	-	Pulse amplitude modulation.
t.d.m.	-	time division multiplex.
r.m.s.	-	root mean square.
P.C.M.	-	Pulse Code Modulation.
C.C.I.T.T.	-	International Consultative Committee on Telephony and Telegraph.
NEXT	-	Near-End Crosstalk.
FEXT	-	Far-End Crosstalk.
MDF	-	Main Distribution Frame.
KP&T	-	Kenya Posts & Telecommunications Corporation.
HDB3	-	High Density Bipolar Code of Order 3.
VF	-	Voice frequency.
RC	-	Raised cosine shaping.
C	-	Cosine shaping.
G	-	Gaussian shaping.
ALBO	-	Automatic Line-Building-Out network.
AMI	-	Alternate Mark Inversion.
DSV	-	Digital Sum Variation.
f_o	-	Nyquist frequency = ½ bit rate = 1024 kHz.
M_N	-	Mean near-end crosstalk noise (dB)
δ_N	-	Standard deviation of NEXT loss at f_o.
M_F	-	Mean far-end cross talk (dB).
δ_F	-	Standard deviation of FEXT loss at f_o.
L_o	-	Cable attenuation in dB, between line repeaters at f_o.

M_{ε}	-	error margin.
$\Delta_N(L_O, RC)$	-	NEXT correction factor for section attenuation other than 20 dB with raised-cosine shaping.
$\Delta_F(s)$	-	FEXT correction factor for channel shaping other than raised cosine.

CHAPTER ONE

INTRODUCTION

The Kenya Posts and Telecommunications Corporation intends to introduce Pulse Code Modulation (PCM) transmission systems in their junction network on a broad scale. However, before these systems are introduced, it is necessary to determine whether or not the existing junction cable network is capable of supporting first-order PCM transmission and with what constraints. The purpose of this project is therefore, to investigate the characteristics of the cable network in order to determine its suitability for use with PCM systems.

The use of PCM on junction cables is arising as a consequence of the rapid rise in traffic over medium distances and the convenience of deloaded audio cables as a transmission medium. In urban areas such as Nairobi, existing cables and duct routes are almost fully utilised and the cost of providing additional similar facilities, involving as it does, major road works in busy city areas, is too high. Carrier transmission is economically viable over longer distances (over 15 km)[3].

Much of the existing line plant, as regards noise and crosstalk, is of a standard which makes PCM mode of creating a multiplex-carrier facility more attractive than traditional methods. In addition, when all parameters including signalling, are considered, the PCM mode appears to compare favourably with its competitors on the basis of cost. For instance, if PCM systems are of 30-channel capacity, the capacity of the junction cables can be increased by a factor of 15 for each two pairs that can be used on the cable. Thus, if crosstalk considerations permit 60% of the cable pairs to be used, the total capacity of the cables can be increased by 9 times, which is a handsome increase.

The application of PCM to junction links is, however, somewhat limited by the considerations that, in general, below a certain distance, the terminal cost makes it difficult to justify in competition with voice transmission on individual cable pairs, and above a certain distance, its demands on bandwidth tend to create a line cost which makes competition with f.d.m. more difficult. It is difficult to define clear demarcations; the choice depends on the nature of the existing line plant and the consequences of switching integration.

The line-transmission problem of PCM is essentially that of transmitting a very high-speed isochronous d.c. telegraph signal, substantially without error, over a cable pair. The PCM system specified by KP & T is the 32-channel system. The channels are time-division multiplexed and encoded by PCM, so that the line signal is a train of 2,048,000 bipolar pulse positions per second. The code recommended by CCITT (14) for this system is the HDB3 code (see chapter 4). Transitorized regenerative amplifiers located in manholes containing loading coils and transmission pairs, detect the pulses and retransmit them along the line. The pulses sent to line from the transmitting terminal will be substantially square pulses in general. After transmission over a length of line, however, the pulses will become modified in shape, owing to the line characteristics, and effects of external interference. This distortion in pulses will cause regenerative amplifiers to give out erroneous outputs resulting in distortion of the original signal.

In the evaluation of the suitability of cables for PCM transmission, the objective is to determine how many PCM systems can be operated in a given cable without exceeding a specified error rate. The error rate, in this case, is the fraction of received pulse positions in which a pulse is present when none was transmitted and vice versa. It has been shown (7) that errors occurring at a rate of 10^{-6} are very difficult to detect by the listener. CCITT recommends a maximum error rate of 10^{-7} for a repeater section.

Though there are several causes of noise on a transmission line, it has been pointed out (7) that for PCM transmission, crosstalk from other PCM systems in the same cable and central office noise (e.g. noise from exchanges) are the only forms of interference that require specific engineering measures to combat. The crosstalk power together with the line attenuation per unit length are shown in Appendices A, B and C to be the major factors in determining the maximum PCM systems in a given cable. The crosstalk noise and cable attenuation increase with the bit-rate on the line. As will be shown in later chapters, measurement of these two parameters at half the bit rate i.e. 1024 Kbits/sec is sufficient to determine the suitability of the cable for PCM transmission. However, a test set which simulates the actual pulse pattern from a 32-channel PCM system, is available and more accurate measurements of crosstalk power and cable attenuation can be made.

In this report, the basic principles of PCM systems as well as the major, sources of interference during PCM transmission over paired junction cables are discussed in chapters two and three to provide the background information required for the investigation. In chapter 4 the operation of regenerative repeaters and the characteristics of cable media as well as the line signal used in the 32-channel PCM transmission are presented since they play an important role in the development of line engineering procedures. These procedures, which are employed in determining the maximum number of PCM systems that can be transmitted in any cable, to satisfy a given error rate objective, are developed in chapter 5. Measurements that are made on selected cables are presented in chapter 6 and are used to predict the cable fill under certain constraints. Appendices A, B and C give the derivations for the relationships between near-end crosstalk, far-end crosstalk, line attenuation and office noise and the maximum number of PCM systems in a cable, while in Appendix D, the power spectral density of the line code used in PCM transmission is analysed. Appendix E gives the general description of the physical cable structure of the junction cables under consideration.

CHAPTER TWO

PRINCIPLES OF PCM

The basic principles of PCM systems are well described in many text-books (2, 3, 5). A qualitative summary of the basic concepts of the system will be given in this chapter, to provide a foundation for the investigation that will follow.

2.1 Time sampling and signal reconstruction.

Consider first a speech waveform such as might be represented by Fig. 2.1. In this diagram, ordinate lines have been drawn at regular time intervals so that they represent the instantaneous values of the speech wave at a sequence of instants. A device which produces a series of very short pulses of current or voltage, so that the amplitudes of the pulses exactly represent the characteristic ordinates in the speech wave (i.e. the output is amplitude - modulated pulses, or p.a.m) is said to be a time - sampling device. Fig. 2.2 indicates one form which such a device may take. The speech-wave source s, is applied to a low-pass filter terminated in a resistance r. By means of appropriate timing pulses, the gate G permits the instantaneous voltage across r to be transferred to the input terminals of the amplifier

A. An appropriate series of time-scale pulses are then obtained across the output load R of A.

In a physically realisable system, the message has the major part of the energy limited to a finite frequency band. The sampling theory, which is based on an ideal limited frequency band, can therefore be applied reasonably safely to a practical case. This theory states that, if the highest frequency component is f_c c/s, the time function of the message cannot assume more than $2f_c$ independent values per second. For this reason, the amplitudes at any set of points t seconds apart where $t = \frac{1}{2f_c}$ completely specify the message. Thus, to transmit an ideal band-limited message of duration T , it is sufficient to send the $2f_cT$ independent values obtained by sampling the instantaneous amplitude of the signal at a regular rate of $2f_c$ samples per second.

It is clear from Fig. 2.2 that if, across points P and Q a second circuit is connected, identical to that shown to the left of PQ and supplied with timing pulses of exactly the same frequency as before but slightly displaced in time, it is possible to transmit speech from this second circuit as time sampled pulses into amplifier A. If, at the receiving end of the circuit, similar gates are employed to direct pulses into the correct low-pass filters, an additional channel will be added to the system. There is clearly no theoretical limit to the number of channels which may be so added,provided that their pulses are uniformly and discretely spaced on the time scale. There will, of course, be a practical limit arising from the limitations of gating techniques.

It can be shown (1, 2, 3) that the spectrum of the sampled message consists of components of the original speech signal plus series of components which fall into discrete families consisting of either the sampling frequency or one of its harmonics, together with upper-and lower-sideband signals located above and below the appropriate harmonic of the sampling frequency. The magnitudes of the harmonics and sideband components diminish with frequency and they are dependent upon the width of the pulse. It is therefore easily seen that if a series of pulses derived by time-sampling can accurately be transmitted to a distant point, the original speech signal may be recovered by a simple process of inserting a low-pass filter to accept only the components that fall within the frequency band of the original speech signal.

2.2 Quantisation

The process of time-sampling described in 2.1, has produced p.a.m. pulses. However, p.a.m is unsuited for transmission over long distances, owing to the difficulty of correcting with sufficient accuracy for the variation of line attenuation and phase shift with frequency. Quite small errors will result in a change in the shape of the received pulse, which in a t.d.m. system, results in cross-talk, thus rendering the system unusable. If however, only certain discrete amplitudes of sample size are permitted, so that, when the message is sampled, the amplitude nearest to the true amplitude is sent; provided that the received signal is not excessively altered by line noise and distortion from that transmitted, it is possible to determine accurately which discrete amplitude the signal is supposed to have. Thus the signal can be reformed, or a new signal created, which has the amplitude originally sent.

Representing the message in this manner, by allowing only certain discrete amplitudes, is called quantising.

Quantisation inherently introduces an initial error in the amplitude of the samples, giving rise to quantisation distortion. This distortion will be discussed more fully in chapter three. One important property of this error is that, whatever the amplitude of the pulse, the magnitude of the error will be determined by the magnitude of the quantum steps, so that the average quantising-error power will be independent of the magnitude of the signal itself. This error power can be made as small as desired by reducing the quantum steps, but since the system may be required to handle speech volumes varying over a large range, the total number of steps required to transmit the largest signals could be too large.

For speech signals, it is found that most of the information is contained in the waveform around the axis, whereas a degree of limiting of the waveform will have no noticeable effect. Also it is found that subjectively, noise is less objectionable for high amplitude signals than for low amplitude ones. Hence, it is desirable to have some form of non-linear quantisation. It may be shown that if the quantum sizes are approximately proportional to the logarithm of the signal level, then it is possible to achieve an approximately constant signal-to-quantisation-error power over a wide range of speech volumes(2). A truly logarithmic characteristic down to zero amplitude would require an infinite number of steps; so that, in practice, it is usual to adopt a characteristic which approximates to logarithmic at large amplitudes, and to linear at low amplitudes. Further details are given in Chapter three and four.

2.3 Coding.

A quantised sample could be sent as a single pulse having, for example, certain possible discrete amplitudes. However, since many sample amplitudes are required (of the order of 100 for speech) it would be difficult to make circuits able to distinguish one from another. On the other hand, it is easy to make a circuit able to determine whether a pulse is present or not. To maintain the required 100 or so levels, the number of such on/off pulses required to represent each sample amplitude must therefore be increased. In general, a code group of n pulses of b possible discrete amplitudes can be used to represent b^n signal amplitudes. For the usual PCM case of on/off pulses, b is 2, and the code is known as binary or base-2 code. Other codes are possible where b = 3 (ternary) or 4 (quaternary) etc; but the emphasis is on binary codes.

Practical systems using non-linear quantisation, need about seven binary digits (i.e. equivalent to 128 levels) to adequately handle the range of speech levels encountered in practice. Such a group of binary digits or bits employed to represent each speech sample is called a character.

Practical systems also need to take account of the difficulties incurred with unrestricted sequences of binary signals which can give rise to long sequences of either marks or spaces (i.e. 1's or 0's). Such signals will not be faithfully transmitted by a transmission medium with a low-frequency cut off (e.g. repeatered cable), and they are also devoid of timing information essential to the regenerator. A number of methods of dealing with this problem are available(2), but the method recommended by CCITT, is that, which in effect, adopts ternary transmission by inverting alternate mark signals (8). These points are dealt with more fully in chapter four.

2.4 Decoding

Decoding of a binary pulse train is obviously the reverse of the encoding process, and it involves generating a pulse which is the linear sum of the pulses in a received code group, each multiplied by its place value in that group (i.e. by 1, b, b^2, b^3 etc). There are basically two different ways of doing this; namely, either by accumulating the received information in a digital memory and decoding all in one step or by decoding each element of information as it arrives, and integrating the code signal in an analogue store. The latter method tends to give large errors, and all practical systems use digital memories.

2.5 Basic advantages and disadvantages of PCM

Sections 2.1, 2.2 and 2.3 have briefly given the three basic processes that are required in the formulation of a PCM signal from an analogue speech signal. It is now appropriate to enumerate briefly the main differences between PCM and f-d-m (or any other analogue system). The many advantages will be seen to justify the current intense interest in PCM.

(a) PCM can accept high levels of line noise or crosstalk or both. It can be shown (3) that a minimum signal/noise ratio (i.e. peak signal power to r.m.s. noise power) of 14 dB for binary transmission (20 dB for ternary) satisfies the maximum error rate of 10^{-7} for speech. This can be compared with a typical minimum repeater-section figure for a mainline f.d.m. system of say 60-70 dB. Thus PCM requires much less signal power, even though the noise power is increased by virtue of the greater bandwidth of the signal.

(b) While the regenerative-type repeaters of PCM are relatively closely spaced (typically 1.8 km for 32-channel system on deloaded audio cables), they are also relatively cheap compared with analogue type repeaters.

(c) As a very generalised statement, PCM compares favourably with its competitors when all aspects including signalling, are considered. When the field of junction-area application embraces city areas, where new cable becomes excessively costly because of the expense incurred with major road disruptions, the cost picture becomes vastly more favourable to PCM - under these conditions, PCM can prove preferable at very short distances.

(d) A transmission facility such as PCM based on digital operation is much more readily suited to the handling of digital-data traffic than one based on analogue principles. The foreseen upsurge in data traffic which could become even more significant given cheap digital-transmission capability, makes this quite an important feature.

(e) The t.d.m. feature of PCM can lead to ready combination of switching and transmission in an integrated network. The multiplexing/demultiplexing functions of f.d.m. do not readily lend themselves to combination with either space or time switching.

(f) In general, a PCM terminal equipment providing a given number of channels has more components than a corresponding f.d.m. system. However f.d.m. systems demand more expensive close-tolerance high-grade components than PCM. Therefore at the present time, there is little difference in the overall cost per channel. As progress is made in digital techniques, the advantage in equipment cost is expected to move in favour of PCM.

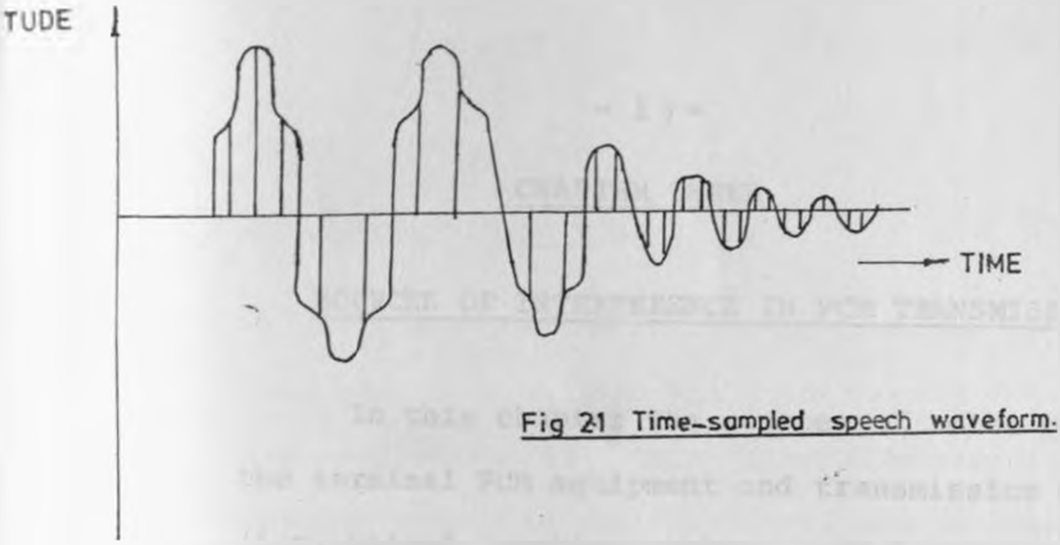


Fig 2.1 Time-sampled speech waveform.

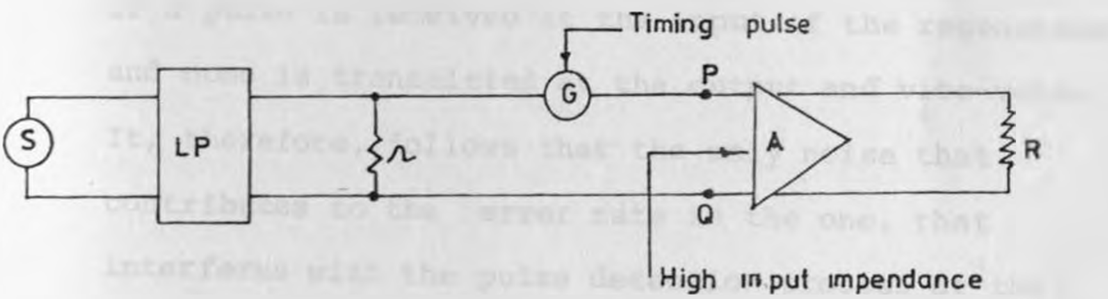


Fig 2.2 Block schematic of time-sampling device

CHAPTER THREE

SOURCES OF INTERFERENCE IN PCM TRANSMISSION

In this chapter the sources of noise both in the terminal PCM equipment and transmission medium (i.e. paired junction cables), are discussed. As stated in the introduction, the system performance is considered satisfactory as long as the error rate per repeater section is 10^{-7} or better. Errors are made at the decision point of the regenerator if a pulse is received at the input of the regenerator and none is transmitted at the output and vice-versa. It, therefore, follows that the only noise that contributes to the error rate is the one, that interferes with the pulse detection process at the input of the regenerator. It will be shown that crosstalk and impulse noise are the major interferences at the decision point of the regenerator, while quantisation noise in the terminal equipment does not contribute to the system error rate and hence does not play any role in cable evaluation. Thermal noise voltages in the cables are too small to enter into line engineering.

3.1 Quantisation noise.

It was pointed out in the previous chapter, that quantisation inherently introduces noise in the sampled signal. This noise increases as the quantum step size increases. It can be shown (1), that in uniform quantisation, with quantum step size, s, the total mean squared quantisation error $\overline{e^2}$ is given by:

$$\overline{e^2} = \frac{s^2}{12} \qquad \text{--------------------} \quad (3.1)$$

This means that if a sample is represented by n binary digits, the signal power-to-quantising distortion power ($S/_D$) is given by

$$S/_D = 6n + 1.8 \text{ dB} \qquad \text{----------------} \quad (3.2)$$

This demonstrates the linear relationship between the number of digits transmitted per sample and the signal-to-distortion ratio in dB. An added binary digit increases the $S/_D$ ratio by 6dB.

However, uniform quantisation is not suitable for message signals since low-level signals which contain most of the information would have more distortion than the strong ones. What is required is an approximately constant signal-to-quantisation-error power over a wide range of speech values. This is achieved by making the quantum sizes approximately proportional to the logarithm of the signal level (except near the origin where a linear characteristic has to be used). In practice all that is required is to pass the signal through a non-linear circuit with a suitable transfer characteristic which compresses the signal and follows it by a uniform encoder. At the remote end, there is a complementary expansion circuit after the decoder. The process is then called companding.

Two companding laws, which satisfy the requirement of constant signal-to-quantisation distortion ratio over a wide range of signals, have been derived (1,2). If the signal amplitude is x, then the compressed signal F(x) is given by:

(i) A-law

$$F(x) = \frac{Ax}{1 + \log A}, \quad 0 \leq x \leq \frac{1}{A}$$

$$= \frac{1+\log(Ax)}{1+\log A}, \quad \frac{1}{A} \leq x \leq 1 \quad \text{-----(3.3)}$$

and (ii) μ-law

$$F(x) = \frac{\log(1 + \mu x)}{\log(1 + \mu)} \quad \text{--------- (3-4)}$$

The parameters A and μ determine the range over which the signal-to-distortion ratio is comparatively constant. The CCITT recommends the A-law for the 30 + 2 PCM system with A = 87.6 giving a companding advantage of 24.1 dB. This is equivalent to the addition of 4 digits, which would be required in the non-companded case to obtain the same error rate. Generally, speech quality will be satisfactory, if the signal power-to-total distortion ratio in the terminal PCM equipment is better than 20 dB (3).

3.2 Thermal noise

Thermal noise in PCM transmission is generated by the resistive components of the transmission medium. The band-limited available noise power P_n which is directly proportional to the bandwidth of the system and the absolute temperature of the medium is given by

$$P_n = kTB_w \text{ watts} \quad \text{--------------------} \quad (3\text{-}5)$$

where k is Boltzman's constant, T is the absolute temperature and Bw is the system bandwidth in hertz. It is apparent that thermal noise power is too small to enter into line engineering (8). For example, the rms thermal noise voltage in a band from zero to two megacycles per second, at a temperature of $300^{o}K$ and resistance 100Ω (typical characteristic impedance of 22-gauge cable at 1024 kHz) is 2 microvolts, while a typical value of the pulse height at the decision point of the regenerator is 0.07 volts, more than 90 dB above the rms thermal noise.

3.3 Crosstalk from other PCM systems

Crosstalk is defined as the disturbance created in one communication circuit by the signals in other communication circuits. In PCM transmission on twisted pairs, crosstalk interference often limits the regenerator spacing or cable capacity. It has been established through extensive measurements (6,7,8) that only two types of crosstalk are of primary importance in PCM transmission. These are near-end crosstalk (NEXT) and far-end crosstalk (FEXT). Other types of indirect crosstalk, although present, are of secondary importance. NEXT and FEXT will now be discussed in detail in the following sections.

3.3.1 Near-End Crosstalk

If a signal is applied to a cable pair termination on the MDF, then any signal induced into other cable pairs of the same cable measured at the

MDF, is termed Near End Crosstalk. i.e. NEXT is the level of induced signal in a cable pair caused by a signal applied to another at the same point as the measured induced signal. NEXT coupling paths are illustrated in Fig. 3.1.

Near-end crosstalk was recognized as a significant problem in one cable systems, (i.e. systems in which both directions of transmission is accommodated in the same cable sheath), during the development and testing of an experimental system (8).

Referring to Fig. 3.1, the high-level outputs of the "disturbing" regenerators are coupled to the inputs of the "disturbed" regenerators via near-end crosstalk paths, the most damaging of which are in the cable near the regenerator casing.

In Appendix A, a theory is developed for predicting the interferring power that arises when n systems operate in a cable. A knowledge of the mean value of this power and of its standard deviation as a combination of n pairs is selected, allows the development of line engineering procedures, as is done in Chapter five. Since NEXT interferring voltage has a Gaussian amplitude distribution (7) error rates for regenerator sections may be calculated using only the rms value of the interference.

It will be noted that variation of NEXT losses with length of regenerator sections and frequency is very erratic if closely observed in any individual instance. However as will be shown in Appendix A, NEXT loss is independent of the length and falls with increasing frequency at about 4.5 dB/octave if the results of many pairs are averaged. Since for the 30 + 2-channel PCM system, the transmitted code (HDB3), has most of the power at half the bit rate i.e. 1024 kHz, the NEXT loss considered in the evaluation of the cables must be at this frequency. (see Appendix D).

3.3.2 Far-End Crosstalk

If a signal is applied to a cable pair termination on the MDF, then any signal induced into other cable pairs(of the same cable) measured some distance from the exchange is termed Far-End Crosstalk, FEXT i.e. FEXT is the level of induced signal in a cable pair caused by a signal applied to another pair at some distance from the measured induced signal. FEXT coupling paths are shown in Fig. 3.2.

It has been pointed out in (7) that far-end crosstalk, is serious only, (in fact it is the only limitation) in a two-cable operation. In the two-cable method of working, one cable is used for all the transmit directions of transmission and a second cable for all the return directions. This method eliminates the effect of NEXT, leaving only FEXT as the limitation, and in general allows more pairs per cable to be used than the single-cable method. However, in the KP & T corporation's junction network under investigation only one cable is laid down for each route and since no provisions have been made for laying down a second cable, the initial evaluation will be based on a one-cable operation.

Nevertheless, the theory of FEXT is developed in Appendix B to provide a comparison between the two methods of operation, if extra cables are laid in future. As in the NEXT case, the statistical properties of the interferring FEXT power when n systems are present in cable, are found. The assumption that the crosstalk voltage has a Gaussian amplitude distribution again allows the estimation of error rates based on interference power. The theory is then applied to the evaluation of two-cable systems and to junctions in chapter five.

Unlike NEXT, FEXT power increases in direct proportion to the regenerator-section length and it usually falls at 6 dB/octave.

3.4 Impulse Noise

Impulse noise is defined as any burst of noise which exceeds the rms noise level by a given magnitude. The primary source of coupling of the impulse noise into a PCM system, is from the central offices.

In a telephone switching office, a variety of devices and circuits create transient and repetitive currents with energy spread over a wide frequency band. The more common sources are relays and switches that interrupt direct or alternating currents, rectifier power supplies and ac power wiring, and wires carrying ringing signals and other plant and test tones Carrier wiring inside the office, e.g. from main distributing frame to carrier equipment bays, can be isolated from these disturbances by shielding and by physical separation. Office noise can also find its way into carrier lines by secondary induction. That is, voice-frequency pairs within the office are exposed to office noise, and a number of these pairs may enter cables which contain carrier pairs in the outside plant. The noise voltages are coupled from the voice-frequency pairs into carrier pairs by crosstalk, chiefly near-end and far-end.

Fig. 3.3 shows a typical PCM line layout in a cable leaving a central office.

Central office noise, coupled to the PCM systems through near-end crosstalk paths, decreases with distance because of its attenuation by the voice-frequency pairs over which it is propagated. Noise coupled through far-end crosstalk paths is also decreased by the voice-frequency pairs, but the crosstalk coupling coefficients themselves increase with length. Thus there is a peak in the noise power at a certain distance.

In Appendix C, some of the properties of office noise are described, and in chapter five, a simple application of these facts is made to arrive at line engineering methods.

3.5 Timing Jitter

Timing jitter is the amount of departure from the correct zero crossing time of marks and spaces- expressed as a fraction of the sampling interval. In PCM transmission, timing jitter is introduced at each repeater and can be shown to accumulate in the repeater chain and may lead to crosstalk and distortion in the reconstructed analogue signal(1).

The sources of timing jitter may be classified as systematic or non-systematic according to whether or not they are related to pulse pattern. Systematic jitter sources yield effects which degrade the pulse train in the same way at all repeaters in a repeater chain. Examples of such sources include intersymbol interference, finite pulse width and clock threshold offset. Non-systematic jitter sources such as mistuning and crosstalk result from timing degradations which are random from repeater to repeater. In a long repeater chain, the total accumulated jitter is dominated by components produced by systematic sources.

It can be shown that the mean-square value of jitter in a long chain of repeaters, increases with the number of repeaters and that jitter is proportional to the timing filter bandwidth (1). Hence high Q-tuned circuits in the timing extractor section of the repeater, will reduce jitter.

Concluding Remark

Having surveyed the various sources of noise in PCM transmission, it is important to note that as long as terminal equipment and repeaters are designed to high standards, then the errors they introduce i.e. quantisation noise and timing jitter can be minimised. As far as cable evaluation is concerned, these errors can be assumed fixed so that the error rate is solely dependent on interference coming from the transmission medium. Therefore, the only impairments to the PCM signal that will be considered in the subsequent analysis, will be near-end crosstalk, far-end crosstalk and noise from central offices as well as attenuation characteristics of the cables and all PCM equipment will be assumed to conform to CCITT standards.

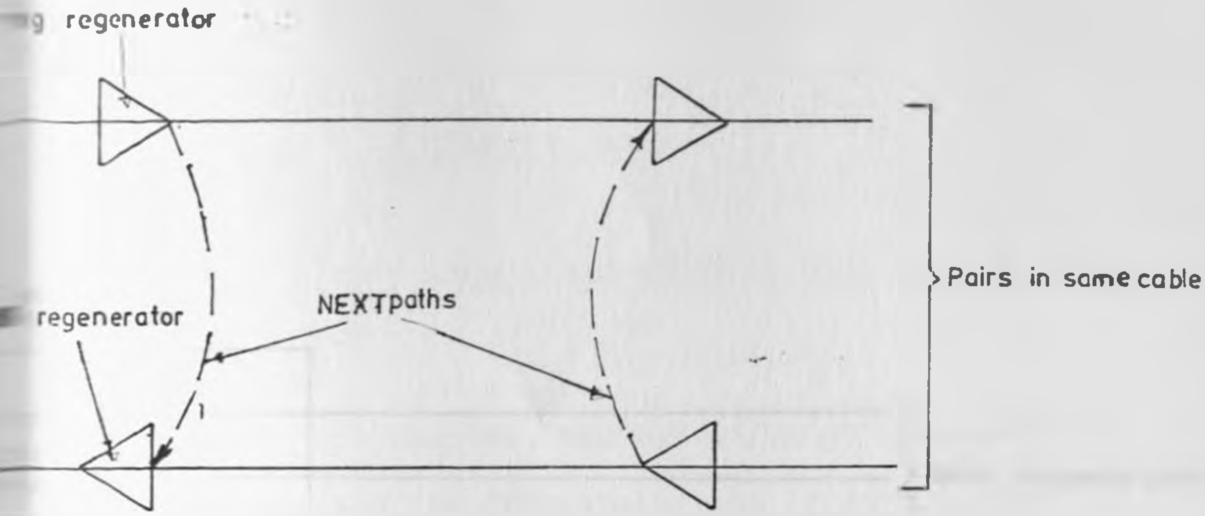


Fig 3·1 Near-end cross-talk paths

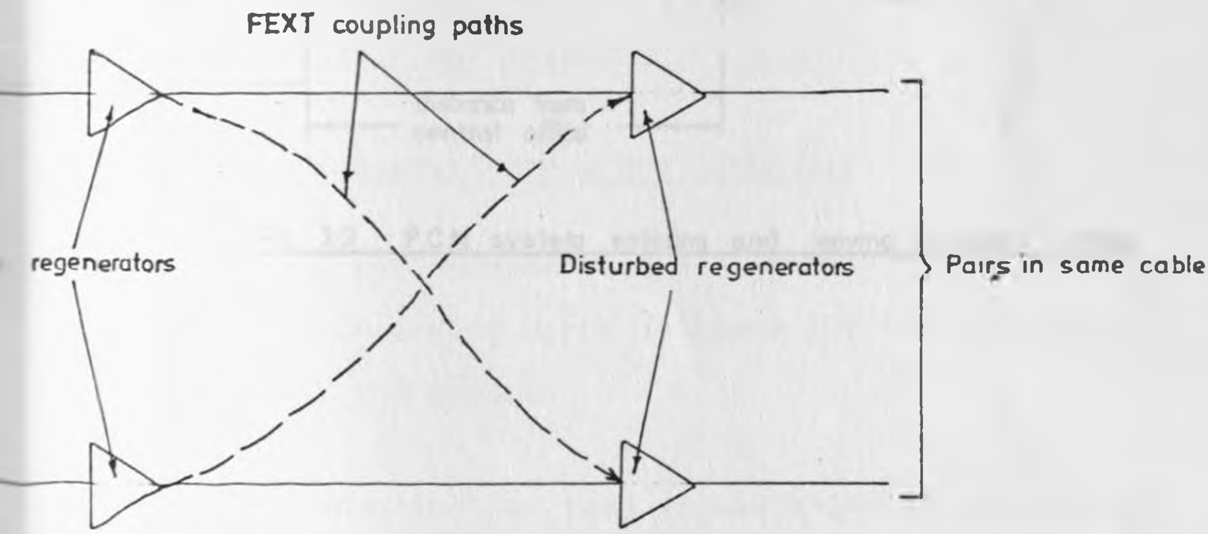


Fig 3·2 FEXT coupling paths

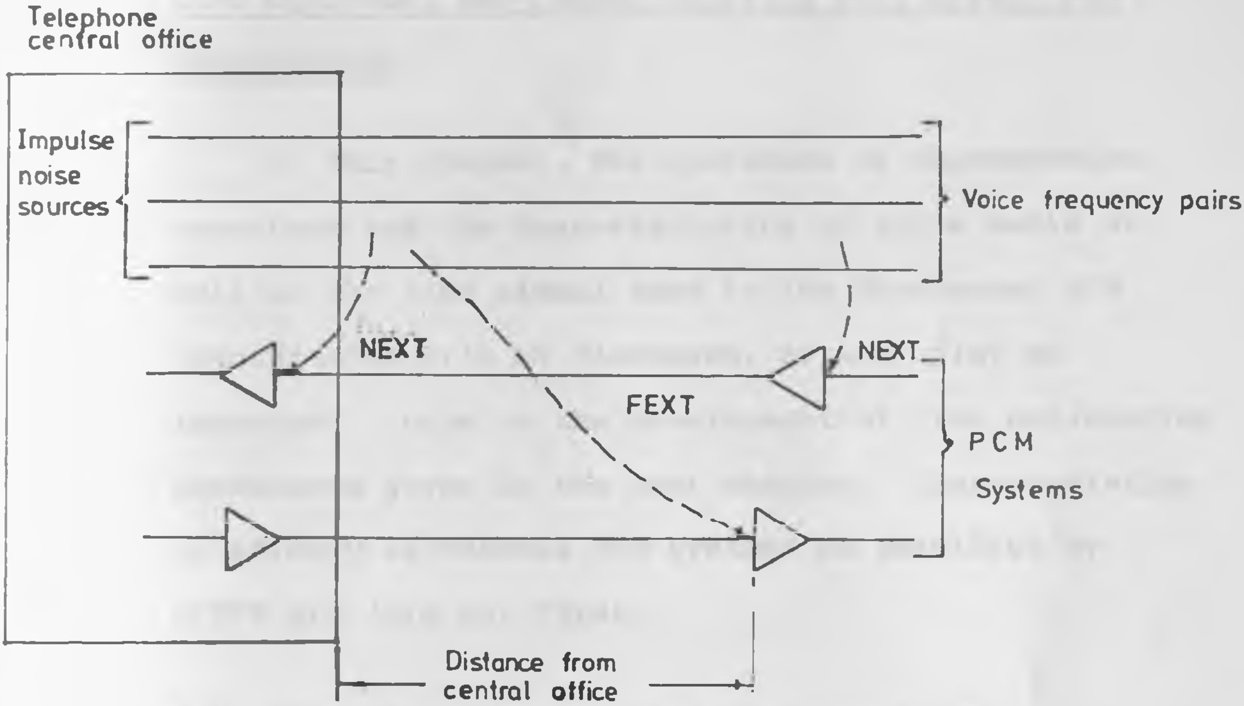


Fig 3·3 PCM system entering and leaving a central office

CHAPTER FOUR

LINE EQUIPMENT AND CHARACTERISTICS THAT AFFECT PCM TRANSMISSION

In this chapter, the operation of regenerative repeaters and the characteristics of cable media as well as the line signal used in the 32-channel PCM transmission will be discussed, as they play an important role in the development of line engineering procedures given in the next chapter. Characteristics of primary 32-channel PCM systems as specified by CCITT are laid out first.

4.1 CCITT Specification for 32-Channel PCM

CCITT recommendation G.732 ammended in 1976 gives the following specifications for the primary 32-channel PCM system:

The encoding law used is the A-law as specified in G.711. The sampling rate is 8000 samples/sec and the number of digits per sample is 8. The nominal bit rate for 32 channels is therefore 8000 x 8 x 32 = 2048 kbit/s, with a tolerance of ±50 parts per million.

The line code to be used is HDB3. The nominal peak voltage of a "mark" is ± 3V ± 10% at 120Ω, and peak voltage of a "space" is 0-0.3V at 120Ω.

Frame structure

The number of bits per channel time slot is 8 numbered from 1 to 8 and the number of channel time slots per frame is 32 numbered from 0 to 31. The channel time slots 1 to 15 and 17 to 31 are assigned to 30 telephone channels, while time slot 0, may or may not contain a frame alignment signal and time slot 16 is assigned to signalling.

4.2 Practical System Block Diagram

A block diagram showing the elemental blocks of a PCM transmission link used to expand installed VF cable capacity is shown in figure 4.1. A span line is composed of a number of repeater sections permanently connected in tandem at repeater apparatus cases mounted in manholes. A "span" is defined as the group of span lines which extend between two office (switching centre) repeater points.

The spacing between regenerative repeaters is important. As will be evident in chapters 5 and 6, it is necessary to remove load coils from these junction cable pairs which are to be used for PCM transmission. It is at these load points that the regenerative repeaters are installed. On a VF line with H-type loading, spacing between load points is normally about 1.83 km. The first load coil out from the exchange on a trunk pair is at half distance or 915 m. This is provident for a regenerative repeater must be installed at this point to increase the pulse level before entering the environment of an exchange area where the levels of impulse noise may be quite high.

4.3 Cable Characteristics

It is impractical to assume that a cable that is known to operate satisfactorily in the voice frequency band will necessarily accept PCM signals. The fault in this assumption is that there are more critical transmission requirements for the high bit rates associated with PCM. A cable pair acts as a low-pass filter to transmitted signals. It has a tendency to pass the lower frequencies quite easily, but rejects the higher-frequency components. In rough figures, the loss per km at voice frequencies (0-4 kHz) is about 1 dB but about 30 dB per km at 1MHz.

Although simple in appearance, a cable pair has complex electrical properties which must be taken into consideration when designing a transmission system, since they determine the transmission characteristics of the cable pair. Figure 4.2, shows a simplified equivalent circuit of a cable pair. The series resistance (R) is the simple ohmic resistance of the conductors. The series inductance (L) is the self inductance of each conductor plus the mutual inductance between the individual conductors. Shunt conductance (G) is the total conductance of the current leakage paths between the conductors, and shunt capacitance (C) is the electrical capacitance between conductors. C_1 is the capacitance between the conductors and ground. These parameters define the attenuation characteristics of the cable and their particular values depend on the physical configuration of the cable, the material of which it is constructed, the frequencies involved and ambient temperature.

It can be shown (1), that at high frequencies, the inductance tends to act as an open circuit while the capacitance approaches a short circuit. The constants contributing to loss, R and G, on the other hand increase with frequency. Thus as frequency increases, the more loss there will be in the cable More specifically, at high frequencies, the attenuation constant $\alpha = \frac{R}{2}\sqrt{\frac{C}{L}} + \frac{1}{2} \quad G\sqrt{\frac{L}{C}}$.

Because of the frequency difference between the voice band and the spectrum used for transmission of PCM pulses, acceptable VF pairs may not be good PCM pairs. Voice usage only requires that a cable pair have consistent properties out to about 4 kHz. PCM requires that the cable perform as predicted out to about 2.5 MHZ. Added to this requirement of a very wide bandwidth are other potential problems such as crosstalk and impulse voice which become more critical at the higher frequencies.

Also, since the cable attenuation increases with cable length, the peak of a transmitted pulse decreases and its base width widens as the cable lengthens (1). In order to successfully detect these pulses, the width must be compressed by means of pulse shaping networks at the receiving end. Pulse shaping networks, also known as, equalizers, not only must compensate for the length of the cable over which the pulse has travelled, but also must take into account the effects of temperature variations on cable constants.

4.4 Regenerative Repeaters

The operations that must be performed in a regenerative repeater, in order to satisfactorily reconstitute a signal that has been distorted during transmission through a band-limited medium are those of reshaping, retiming and regenerating.

4.4.1 Reshaping (Equalization)

Reshaping involves the provision of equalization at the input to the repeater; so that the combined overall frequency characteristic of the transmission medium plus the equalizer is such as to produce, at the input to the regenerator, a pulse train, to which it is possible to allocate a slicing level for the purpose of making a decision as to whether a mark or space is present. It can be shown that a gradual roll off rather than a sharp cut off, amplitude characteristic is desirable for reducing interdigit interference (1). Such interdigit interference also arises from pulse echoes produced by ripples in the overall amplitude or phase frequency characteristic or both.

There are three common types of channel shaping which are used to minimize intersymbol interference mainly,raised-cosine (100% roll-off), cosine and Gaussian shaping. In Appendix A and B, it is shown that near-end and far-end crosstalk interference from other PCM systems, at the input of the regenerator depend on a number of factors including the type of channel shaping used. It is further indicated that for any given code, repeaters using raised cosine shaping suffer the lowest crosstalk interference. Raised-cosine channel shaping will therefore be assumed throughout this investigation. Corrections factors, if other types of shaping are used, will also be given provided (see next Chapter).

4.4.2 Retiming

After equalization, the pulse train in the repeater is in a form suitable for regeneration. The timing circuits in the repeater control the regeneration process by providing a clock signal to (a) sample the equalized pulse train near the centre of each pulse (b) maintain proper pulse spacing at the regenerator output (c) assure correct pulse width.

The timing signal is extracted from the pulse train after equalization and amplification. It is important to note that no discrete signalling rate component is present in the spectrum of the transmitted code (HDB3). With HDB3 code, the spectrum falls to zero at the bit rate, but there is a significant continuous component at half the bit rate (see Appendix D). The timing waveform may therefore be derived by rectifying the band-limited signal and applying the train of pulses produced by the cusps of the rectified waveform to a narrow band filter centred on the bit frequency. The extracted component must be amplified in a high Q-factor circuit or else the timing component will decay during long periods of spaces. As will be seen in section 4.5, special codes which do not permit long periods of spaces have to be used so as to avoid loss of the timing component.

4.4.3 Regeneration

The timing component after amplification, is limited to produce an approximate square wave at the signalling rate. The resulting signal controls a clock-pulse generator which yields narrow positive and negative clock pulses at the zero crossings of the square wave. The positive going pulses are used to gate the incoming equalized pulse train to the regenerator, which gives out a pulse output if the equalized pulse amplitude exceeds a present level. The negative-going clock pulse, similarly turns off the regenerator, thus controlling the width of the regenerated pulses.

4.4.4 Automatic Line-Building-Out Network

Since repeater sections vary in length, pulse trains will have varying distortions after travelling through different sections. The automatic line-building-out network (ALBO) automatically controls variations in repeater spacing. The peak of an incoming pulse is detected by a peak detector which then feeds a control signal into a control element so that the same peak value is maintained at all times, regardless of cable length. As the cable length varies, the ALBO readjusts to maintain what appears to be a constant length going into the equalizer of the repeater.

A simplified block diagram of a regenerative repeater is given in Figure 4.3. Waveforms coming out of each component are also indicated.

4.5 Line Coding

The line coding recommended by CCITT for 32-channel PCM transmission is the high-density bipolar code of order 3 (HDB3). This is a bipolar code in which 4 consecutive zeros are replaced by one of the following sequences:

BOOV or OOOV

where B is a positive or negative pulse generated using bipolar rule

and V is a pulse inserted or inverted in the original sequence and violates the bipolar rule i.e. V and B have the same polarity.

The choice of these sequences is carried out in such a way that the number of B pulses between consecutive V pulses is always odd. V pulses are therefore alternating in polarity and consequently the digital sum variation of the code (which is a measure of the peak to peak distortion when the pulse train goes through an ac network), is only 2.

Before explaining why the HDB3 code is chosen in preference to several other codes, it is necessary to describe briefly the formation of the bipolar code from which HDB3 code is derived. The bipolar or alternate mark inversion (AMl) code, is generated from the simple binary code by

(a) representing each binary zero with a space.
(b) representing each binary one with a pulse or mark.
(c) alternating the polarity of adjacent pulses.

The simple binary code is unsuitable for line PCM transmission because it contains insufficient timing information required in the repeater and it has a DC component and large low-frequency spectral energy which are both objectionable if the line is to be AC coupled for power feeding purposes.

On the other hand, the bipolar code and consequently HDB3 code both have very small power spectra in the low-frequency range and zero power at zero frequency (see Appendix D). Hence the power carried in these codes will not be affected by transformers or capacitors that are used for AC power coupling to repeaters along the line. The second important advantage is that the bipolar and HDB3 codes have sufficient timing content so that clock information can be extracted within the regenerative repeater.

However, though the bipolar code is easier to implement and has the lowest possible digital sum variation (DSV) of unity, its main disadvantage is that zeros in the binary input data are not encoded. It is therefore possible to have long strings of zeros from which timing information cannot be derived. With HDB3 code, any string of 4 or more zeros in the bipolar code, is replaced with a present pattern thereby avoiding loss of timing information. Though there are several other codes which extract long strings of zeros in the bipolar stream (e.g. PST, B6ZS etc), HDB3 is preferred because it has less average power and hence less crosstalk interference. In addition, the code allows transmission errors to be monitored without reference to the transmitted information. Errors produce violations of the bipolar polarity alternation rule or "bipolar violations" These can easily be detected, except when errors occur successively in even numbers and with alternating polarities.

Decoding procedures basically consist of removal of the filling sequences,and rectification of the resulting stream. A more detailed acccount of codes in common use is given in refs (16, 17 and 18).

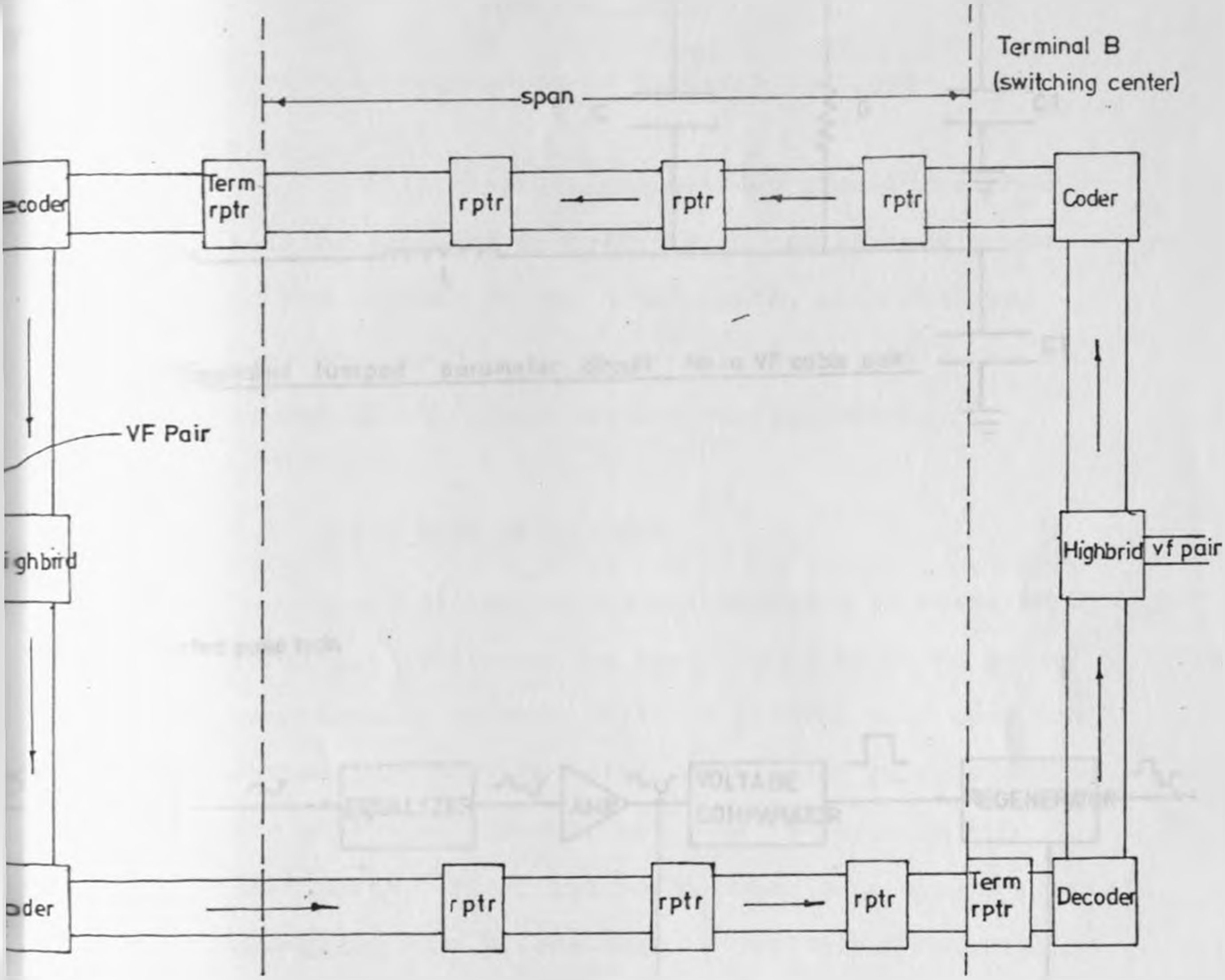


Fig 4.1 Simplified functional block diagram of a PCM link used to expand Capacity of existing Vf cable (one vf pair shown)

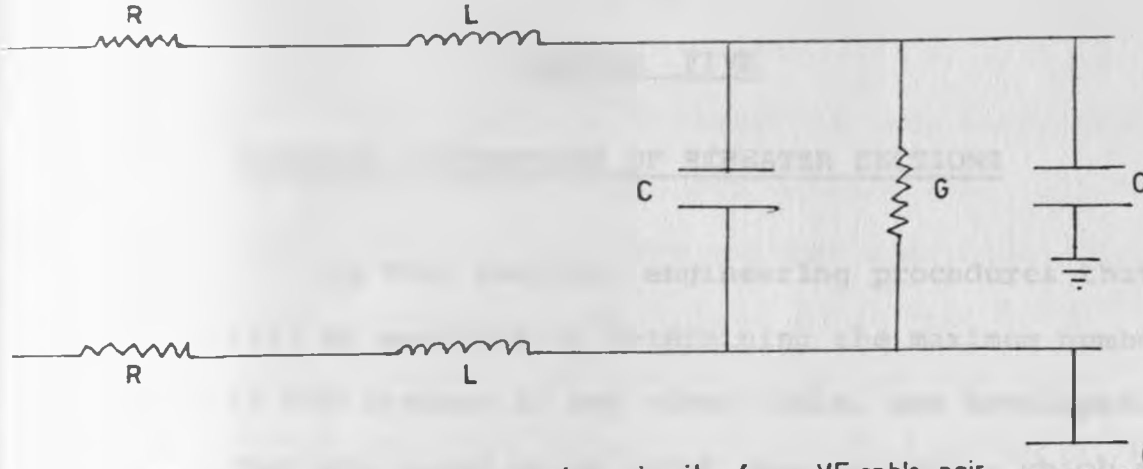


Fig4.2 Equivalent lumped parameter circuit for a VF cable pair

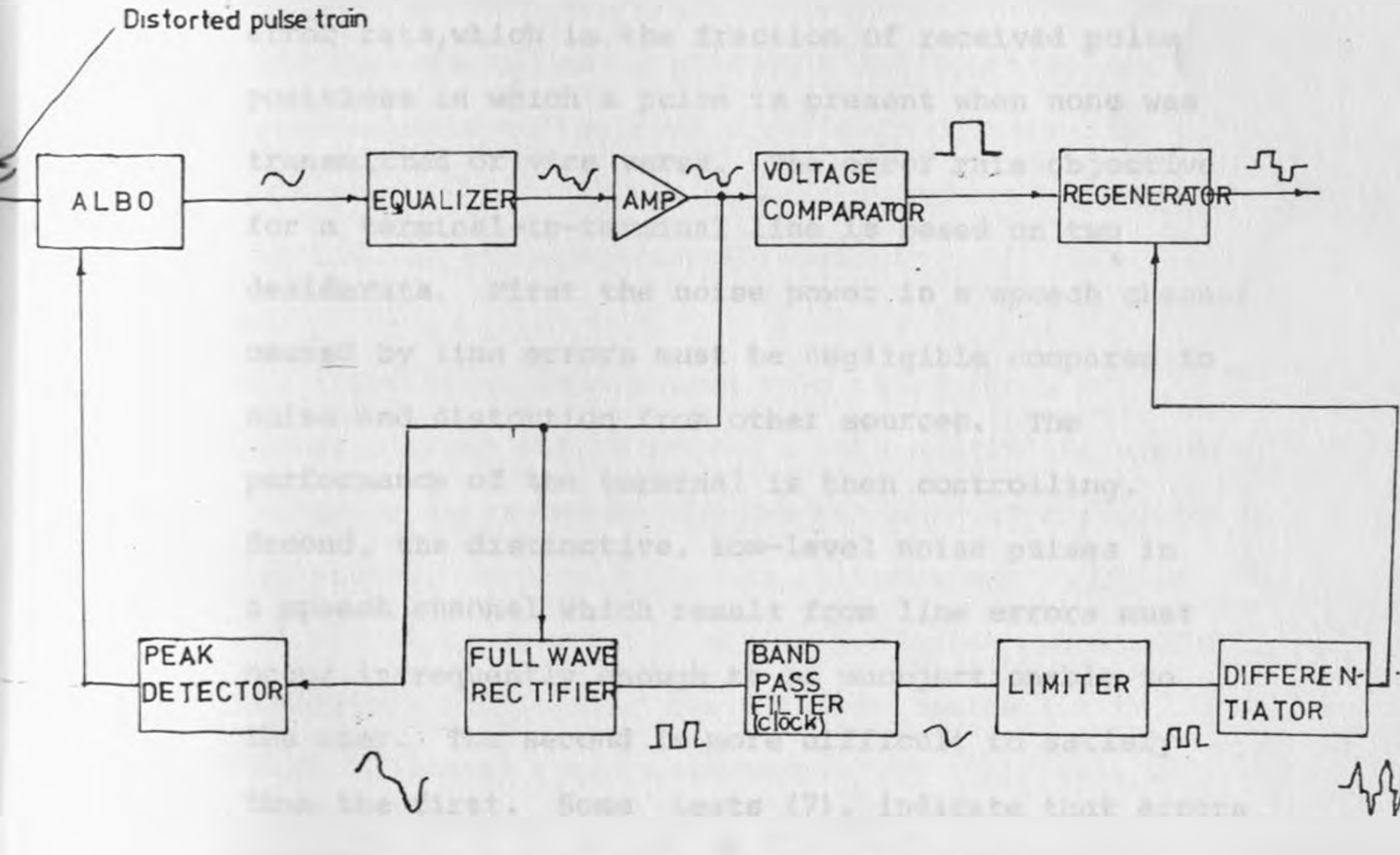


Fig 4.3 Block Diagram of a regenerative repeater

CHAPTER FIVE

DETAILED ENGINEERING OF REPEATER SECTIONS

In this chapter, engineering procedures that will be employed in determining the maximum number of PCM systems in any given cable, are developed. They are based on an error rate objective which must be met in all cable lay-out configurations.

5.1 Error Rate Objectives

The fidelity of PCM transmission is measured by error rate,which is the fraction of received pulse positions in which a pulse is present when none was transmitted or vice versa. The error rate objective for a terminal-to-terminal line is based on two desiderata. First the noise power in a speech channel caused by line errors must be negligible compared to noise and distortion from other sources. The performance of the terminal is then controlling. Second, the distinctive, low-level noise pulses in a speech channel which result from line errors must occur infrequently enough to be unobjectionable to the user. The second is more difficult to satisfy than the first. Some tests (7), indicate that errors occurring at a rate of 10^{-6} are very difficult to detect by listening, and this rate is usually taken as the system objective.

However, it is necessary to have satisfactory transmission over the lines of each span (see section 4.2) To allow the average three span lines in a system to have the maximum error rate of 10^{-6}, the span error rate objective is set at 3×10^{-7}. Since most span lines will have substantially lower error rates, the system objective will almost always be met, even in long systems.

In engineering of a repeater section one has to take account of errors due to crosstalk and office noise so as to produce a sufficiently small contribution to the span error rate objective of 3×10^{-7}, while achieving nearly maximum economy in the use of regenerative repeaters.

The error rate caused by a combination of crosstalk and office noise is very nearly the sum of error rates caused by the two kinds of interference separately (Appendix C). It is therefore possible to set an error rate objective for noise and one for crosstalk and to find the separate design limits, that will meet these objectives. An error rate of 10^{-7} is therefore assigned to office noise in one section so that the two sections in a span exposed to office noise contribute an error rate of 2×10^{-7}.

The remaining error rate of 10^{-7} is then assigned to crosstalk. Because of the extreme sensitivity of error rate to small differences in crosstalk loss or section loss, error rates due to crosstalk will vary radically from one section to the next. The section error rate objective of 10^{-7} is set for the worst one per cent repeater. This means that in a five-section span, about 95% of the lines will meet the span objective, which is a reasonable performance level.

5.2 <u>Data Required for Repeater-Section Design</u>

As will be evident in the remainder of this chapter, there are two types of data that have to be known before a complete evaluation of the suitability of junction cables for PCM transmission, can be made.

Firstly, the topographic layout of the cables within the plant must be known. This data is necessary because different cable layouts will suffer varying amounts of crosstalk interference and office-noise. The KP and T network under investigation has three different layouts of cables namely,

(a) terminal sections which are near and/or entering a central office (e.g. exchange).

(b) normal intermediate sections, which are not adjacent to any terminal station and do not have any branching points.

and (c) intermediate sections with branching.

Constraints governing the maximum number of systems for each of the three confugirations will be given in sections 5.3 to 5.6.

The second type of data required is the electrical characteristics and make-up features of the cables. The desired electrical properties include the characteristic impedance and the mean values and standard deviations at half the bit rate (1024 kHz) of the: attenuation per km, near-end crosstalk attenuation and far-end crosstalk loss. Knowledge of the formation features of each cable offers the possibility of choosing the pairs in order to minimize the harmful effects of crosstalk, and hence to quite fully exploit the cable.

While topographic layout and cable formation data are available in the KP & T network records, the electrical characteristics at the desired frequency have to be termined by measurement. Details of measurement procedures and results will be given in Chapter six.

5.3 One Cable Operation

5.3.1 Relationship Between Near-End Crosstalk and Maximum Number of Systems in a Cable

In one cable systems, both directions of transmission are accommodated in one cable sheath. It was mentioned in subsection 3.3.1 that only near-end crosstalk effects need to be considered in one cable systems. The number of systems, n, installed in a given cable, or the repeater section loss or both, must be limited to satisfy the error rate objective of section 5.1. Since in the KP & T network, it is intended to position repeaters in the existing manholes, the repeater section losses are fixed so that the only variable parameter is the number of systems. The quantitative limits are obtained as follows:

In Appendix A, a quantitative relationship governing the average near-end crosstalk (NEXT), M_N, repeater section attenuation or loss, L_o and the maximum number of PCM systems n in a given cable is established and is given by (equation A.20),

$$M_N - L_o - M_\varepsilon - \Delta_N(L_o, RC) \geq 13.46 + \Big[I(\sigma_N, n) + 2.33\sigma_\phi(\sigma_N, n) \Big] dB \text{ ------------------------ } (5.1)$$

Where $\Delta_N(L_O,RC)$ is a correction factor for repeater section losses other than 20 dB and M_ε is the error margin allocated for other noise sources such as intersymbol interference or misequalization, thermal noise, timing jitter and incomplete knowledge of crosstalk data. σ_N is the standard deviation of M_N. The statistical quantities $I(\sigma_N,n)$ and $\sigma_\phi(\sigma_N,n)$ are given by (equations A.9, and A.10),

$$I(\sigma_N,n) = 5\left[\log(n^3\exp\{0.053\sigma_N^2\}) - \log(\exp\{0.053\sigma_N^2\} + n-1\right] \quad \text{------------------------(5.2)}$$

and $$\sigma_\phi(\sigma_N,n) = 6.593\left[\log(\exp\{0.053\sigma_N^2\} + n-1) - \log n\right]^{\frac{1}{2}} \quad \text{---------------(5.3)}$$

In deriving equation (5.1), it is assumed that (a) the interference NEXT power from n systems has a Gaussian distribution, (b) the type of channel shaping that is least vulnerable to crosstalk interference i.e raised-cosine shaping, is used (see subsection 4.4.1) and (c) the line code is HDB3. The equation quarantees that 99% of the repeater sections will have error rates equal or better than 10^{-7}. The equation is centred on a section loss of 20 dB, because KP & T have indicated that the repeater section loss in their junction cables, varies between 5 and 35 dB at

1 MHz so that 20 dB is the median value.

Since extensive measurements (6, 7, 11, 12) have shown that the standard deviation, σ_N, of NEXT loss varies between 5 and 14, dB we expect cables under investigation to have similar behaviour. It is therefore useful to have plots of NEXT loss versus maximum number of systems for various standard deviations. This was done by computing (with the aid of a computer program) the RHS of equation (5.1) for values of σ_N = 5, 6, 7, 8, 9, 10, 11, 12 13, and 14 and values of n ranging from 2 to 1000 systems using equations (5.2), and (5.3). By setting $M_\varepsilon = 0$ and $\Delta_N(L_o,RC) = 0$ for L_o = 20 dB, curves of $M_N - L_o$ versus n for the above values of σ_N were drawn and are given in figure 5.1. Any desired error margin M_ε (which is usually between 6 and 12 dB) and correction $\Delta_N(L_o,RC)$ for any section loss other than 20 dB can easily be added to the results. The correction curve for section losses other than 20 dB is given in figure 5.2 which was obtained using equation (A.21) in Appendix A, and data from Ref. (6).

It should be noted that plots in figures 5.1 and 5.2 are valid for raised-cosine channel shaping only. For the other most commonly used type of channel shaping, (Gaussian shaping), the NEXT interference increase over that in raised cosine shaping varies from 0 dB for section loss of 5 dB to 0.5 dB for section loss of 35 dB. Thus for all practical purposes, effects of changes in channel shaping can be taken care of by increasing the error margin M_ε by about 0.5 dB.

The plots in Fig. 5.1 give the upper bound on both section loss L_o, and number of systems n for a measured mean NEXT loss M_N and its standard deviation σ_N. Since the repeater section lengths will be fixed by positions of existing manholes, repeater section losses will similarly be fixed. Therefore, the number of systems that may be operated in a repeater section will have to be equal to or less than the upper bound given by the plots, if the error rate objective is not to be violated.

5.3.2 Selection of Cables and Pairs

It would be convenient if one could assign systems to cable pairs as they are available along spans, without selecting individual pairs within the cable sections. Such assignment is possible with voice-frequency circuits, but PCM systems need more careful consideration. For one-cable operations, the two directions of transmission must be assigned to groups of pairs with a satisfactory range of near-end crosstalk coupling losses. More exactly, one would like to use groups of pairs within a cable which have a high mean loss M_N and a small standard deviation σ_N, so that the maximum number of PCM systems possible is high, as is evident in fig. 5.1.

The majority of the KP & T cables under investigation belong to the type described in Appendix E. i.e. they consist of star-quads arranged in concentric layers. Depending on the selection criteria that may be used, the following five different modes of cable filling can be distinguished for the star-quad multi-layer cables.

Mode A: This is the choice in which opposite directions of transmission are in separate and non-adjacent layers.

Mode B: Opposite directions of transmission are in separate and adjacent layers.

Mode C: Opposite directions of transmission are in the same layer and in alternate quads.

Mode D: Opposite direction of transmission are in the same layer and in quads separated by one non-occupied quad.

Mode E: Opposite directions of transmission are in separate quads without any restriction.

Fig. 5.2 shows a typical cable cross-section with the five modes.

Each of the above choices has merits and drawbacks. Mode A has excellent crosstalk performance but at the expense of cable fill-up (50% maximum).

Modes B and C permit good fill-up (up to 100%) but with medium crosstalk performance.

Mode D is limited in fill-up (up to 50%) and has poor crosstalk performance for star-quad cables.

With Mode E, crosstalk performance is not quite good, especially if considering the wide spread, but there is a very considerable advantage in that no previous knowledge of cable characteristics is required.

In Chapter 6, details of measurements made for the various modes will be given and the resulting number of PCM systems for each mode deduced. It will then be easy to tell which cables are suitable for PCM transmission and under what modes.

5.4 Two-Cable Operation

In two-cable systems, one cable is used for all the transmit directions of transmission and a second cable is used for all the return directions. As it was pointed out in sub-section 3.3.2, far-end cross-talk is the only limitation in two-cable operation. This permits, in general, more pairs per cable to be used for PCM transmission, than in the single-cable method.

As in the one-cable method, the number of systems n, installed in a given cable, must be limited to satisfy the error-rate objective of section 5.1. Unlike, the NEXT case, the number of systems that can be installed in a cable, is dependent on far-end crosstalk and channel-shaping only, and does not depend on repeater section loss. In Appendix B, the relationship governing the mean FEXT loss, M_F, its standard deviation σ_F and the number of systems, n,toensure a worst error-rate of 10^{-7} for 99% of the repeater sections, has been found to be (equation B.5, Appendix B).

$$M_F-\Delta_F(S)-M_\varepsilon \geq 13.76 + \Big[I(\sigma_F,n) + 2.33\sigma_\phi(\sigma_F,n) \Big] \text{ ---------------- } \quad (5.4)$$

where the error margin M_ε and statistical quantities $I(\sigma_F,n)$ and $\sigma_\phi(\sigma_F,n)$ are as defined in subsection 5.3.1 with σ_N replaced by σ_F. In equation (5.4), the correction factor $\Delta_F(S)$ is zero for raised-cosine shaping and equal to - 0.3 dB for Gaussian shaping.

Examining equations (5.1) for NEXT loss and (5.4) for FEXT loss, it is obvious that curves in fig. 5.1 can be used to determine the maximum number of systems, n, if M_F and σ_F are known. This is done by substituting M_F - 0.3 for $M_N - L_o$ on the ordinate axis.

As it was mentioned in subsection 3.3.2, the KP & T junction network under investigation will operate on a one-cable basis initially (as there is no second cable) so that this analysis based on far-end crosstalk is included only for providing a more complete treatment and to be used in future if extra cables are provided.

5.5 System Junctions

In an extensive area with PCM systems connecting a number of central offices, junctions such as that shown in figure 5.3 occur. In the figure, systems connect office A with office C, and C with B, but for simplicity no through systems from A to B are included. Only junctions in one-cable systems will be considered.

With reference to fig. 5.3, the dimensioning criteria varies according to the length of L_3, of the common section and is, in any case, checked by the conditions of regenerators atC. If L_3 is very short (about 10 metres), AC and BC can be considered as completely separate links, therefore each must be independently dimensioned as per section 5.3.

If the length of L_3 is some hundred metres or more then sections A-C and B-C can be designed as if the number of systems for each of them were $n_1 + n_2$, where

n_1 = number of systems from office A

and n_2 = number of systems from office B.

This approach is justified by the fact that near-end crosstalk reaches an asymptotic value within approximately the first 6-7 dB of attenuation with transmission at 1024 kHz.

5.6 PCM Systems Near Central Offices

Repeater sections that terminate in central offices, and others in which switched telephone pairs share a cable sheath with PCM lines, are subject to office noise interference (see section 3.3.) The fundamental information for engineering these sections is in Appendix C. As shown there, the signal-to-noise ratio, and hence error rate, of a regenerator repeater is determined by the difference between loss in the path by which the noise reaches the repeater input and the loss in the signal path. An error rate of 10^{-7} is attained if this difference is at least 53 dB under very severe, noise conditions so that this is adopted as the objective. The three most common configurations of cable layout near central offices will now be analysed.

5.6.1. PCM systems entering an office

Fig. 5.4 shows the incoming PCM pair in a simple entrance section. For the office repeater, the signal path loss is L_1, the pair loss between the nearest outside repeater and the main frame. The noise path loss, for the near-end crosstalk coupling path A, may be taken as 76.8 dB (see Appendix C). Therefore L_1 must be given by:

$$76.8 - L_1 \geq 52 \text{ or } L_1 \leq 24.8 \text{ dB.}$$

For the first outside repeater, the signal path loss is L_2, and the noise path loss is $L_1 + 76.8$, for the near-end path B. It follows that

$$L_1 + 76.8 - L_2 \geq 52 \text{ or } L_2 \leq 24.8 + L_1.$$

Thus the length of the terminal sections in this case is half the length of the previous section, which is designed as per section 5.3.

5.6.2 PCM Systems Not Entering the Office

Fig. 5.5 shows an arrangement in which PCM systems do not enter the office, but are exposed to disturbing pairs which do enter it. The repeater operating with signal loss L_4 has a total noise path loss for path C, of $L_6 + M_F$, where M_F is the mean equal-level far-end crosstalk coupling loss at 1024 kHz for the cable section between points X and Y. (The distance from X to Y in metres may be substituted for L in equation (C.1) - Appendix C, to approximate M_F). The losses L_4 and L_6 must satisfy the relation

$$L_6 + M_F - L_4 \geq 52.$$

For near-end path D, the noise path loss L_7 + 76.8, for the repeater that operates with signal loss L_5, so that the required relation is:

$$L_7 + 76.8 - L_5 \geq 52.$$

5.6.3 PCM System cross-connected at office MDF

A third arrangement is that of fig. 5.6, in which the PCM systems are cross-connected at the main frame, for flexibility in installing a terminal at the office later. For noise path E,

$$L_9 + M_F - (L_9+L_{10}) \geq 52 \text{ or}$$

$M_F - L_{10} \geq 52$; here M_F is the far-end coupling loss for the entire L_9 section. For near-end path F,

$$L_{10} + 76.8 - L_{11} \geq 52 \text{ or } L_{11} \leq 23 + L_{10}.$$

5.7 Effects of Line Impedance, Reflections and Temperature Variations.

The repeater output and input impedances are not perfectly matched to the characteristic impedance of the cable pairs (120Ω) over the band of frequencies needed for pulse transmission. As a result, reflections of pulses may travel back from a repeater input to the output circuit of the previous repeater and interfere with the regeneration of pulses there, increasing the error rate. It has been shown (7) that degradation due to reflections is serious only if the length of intermediate repeater sections is reduced below certain limits, a situation that may arise adjacent to central offices (see section 5.6).

The problem concerning reflections is very complex and no simple rules have so far been set. The minimum section loss that may be used to avoid reflections, varies with repeater design. In ref. (7), the error rate is not seriously affected as long as the repeaters are separated by more than 9 dB of cable loss. Central office repeaters with section loss less than this value are equipped with fixed pads of 100 ohms impedance and 3 dB loss, at both input and output. The pads also reduce reflections from cable plugs and terminating cable discontinuities at the cable vault. Other PCM systems manufacturers such as Telettra advise a minimum section loss of 12 dB for cable characteristic impedance that fall outside the range 120 $\pm$ 10 ohms. The lower limit of section loss will therefore depend on the recommendations of the manufacturer of the repeaters that will eventually be installed. For this investigation, terminal sections with losses less than 12 dB will be pointed out as they may have to be lengthened for satisfactory performance.

The primary constants of a cable, and consequently the characteristics impedance and attenuation, vary with temperature. However the variation are very small (on the average the temperature coefficient of characteristic impedance and attenuation is 0.02%/^{o}C (7) at high frequencies and are therefore of little or no importance. The variations are slightly lower for paper-insulated cables than for polyethlene-insulated cables.

FIG 5·1 ESTIMATE OF MAXIMUM NUMBER OF PCM SYSTEMS AS A FUNCTION OF MEAN NEXT AND FEXT FOR VARIOUS STANDARD DEVIATIONS

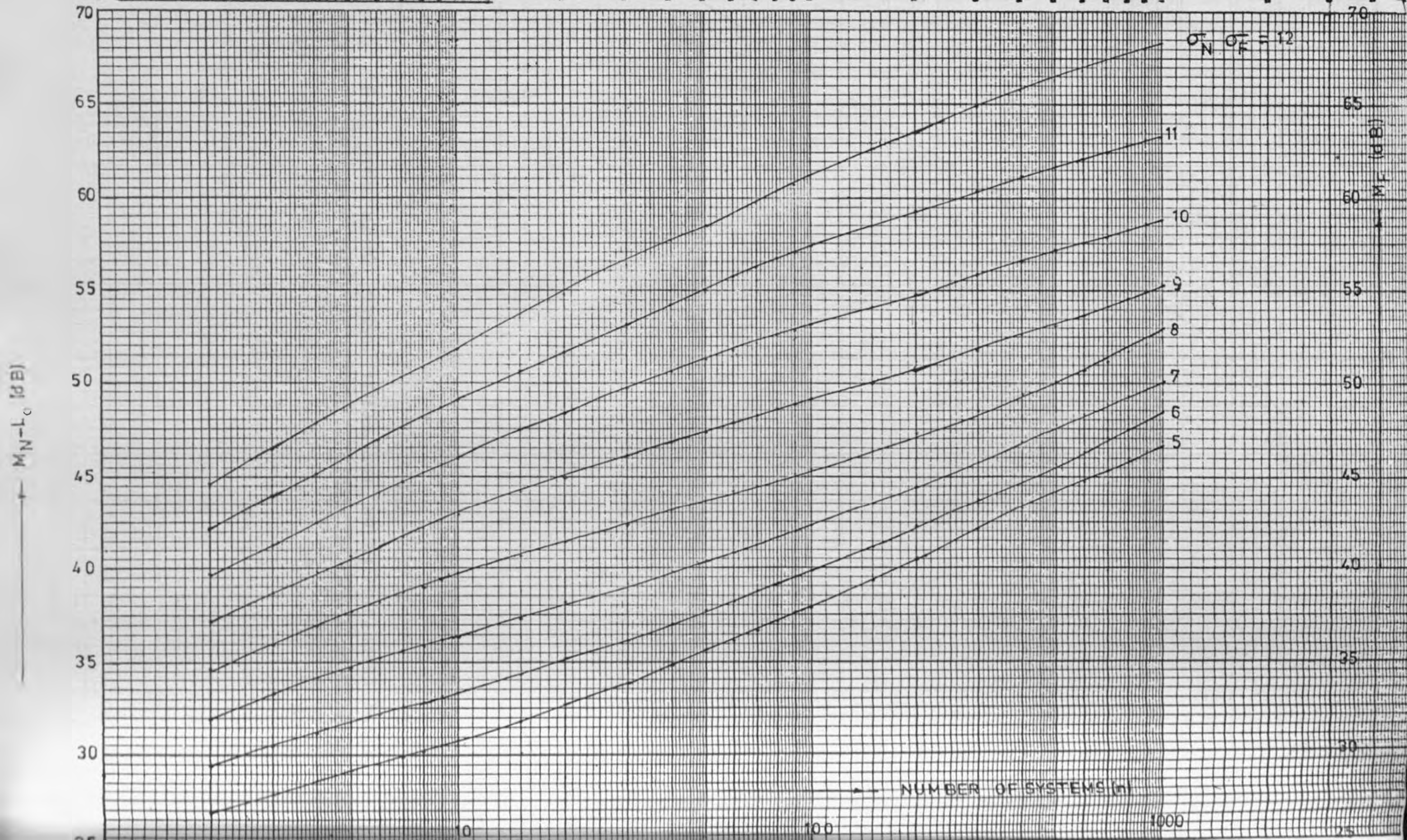
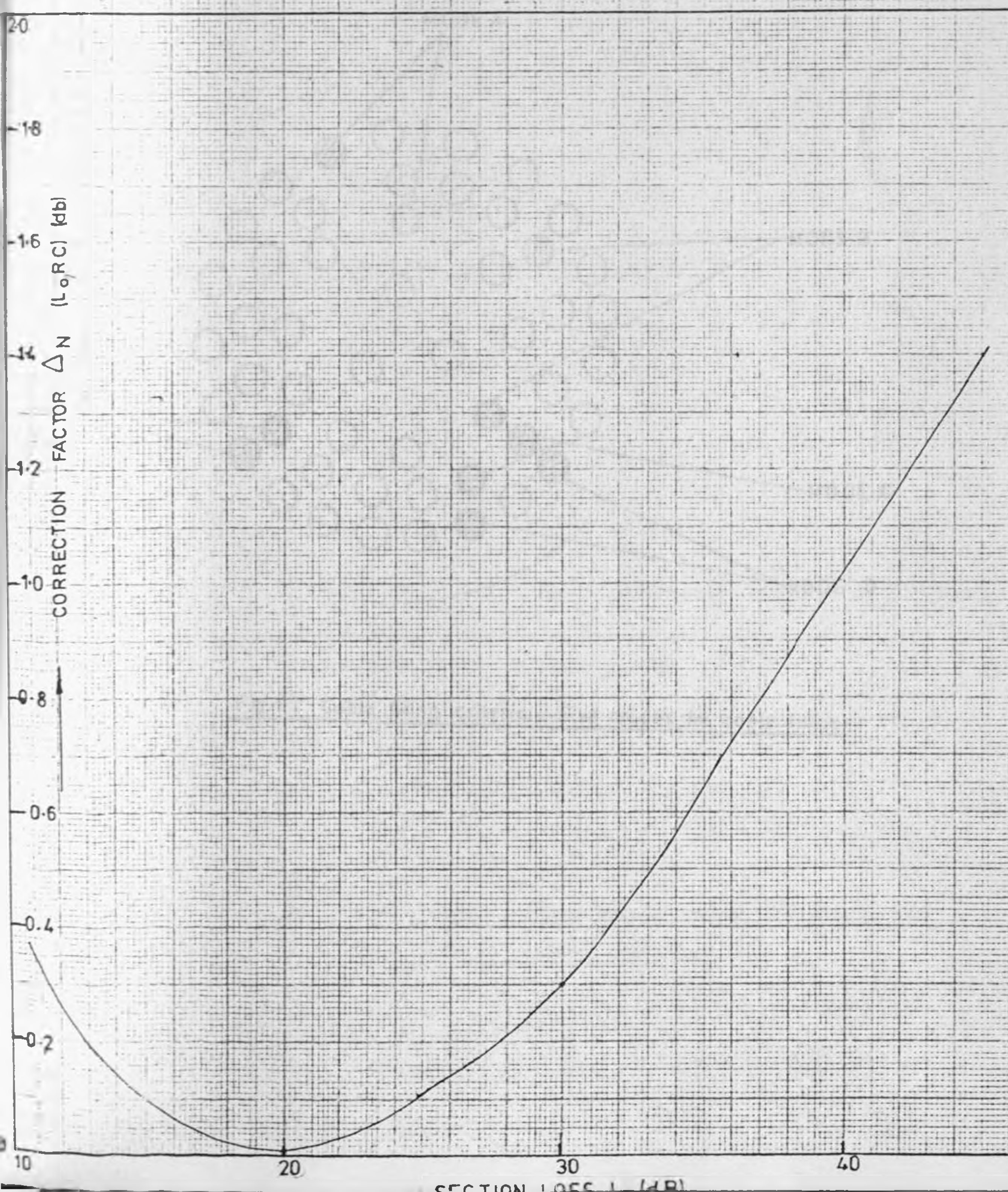


FIG 5·2 CORRECTION FACTOR FOR SECTION LOSSES OTHER THAN 20 dB



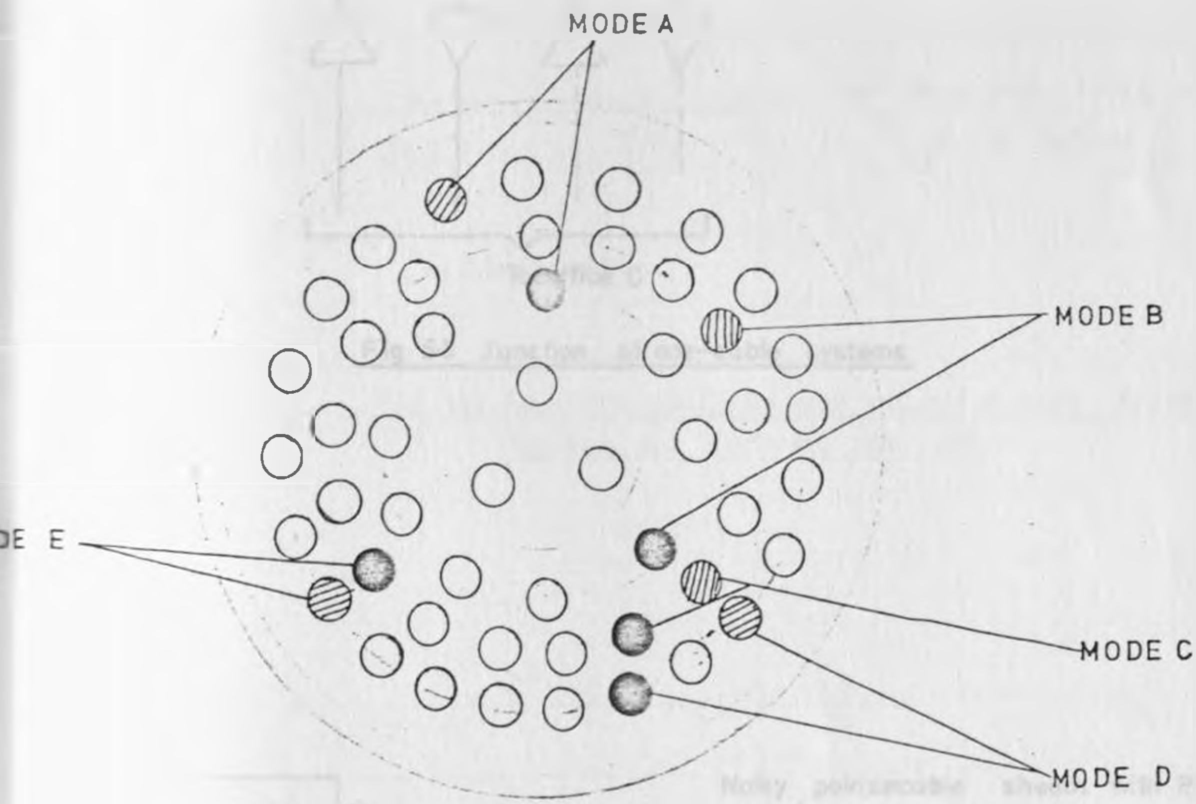


Fig 5·2 Cable filling criteria — Five modes for cable filling.

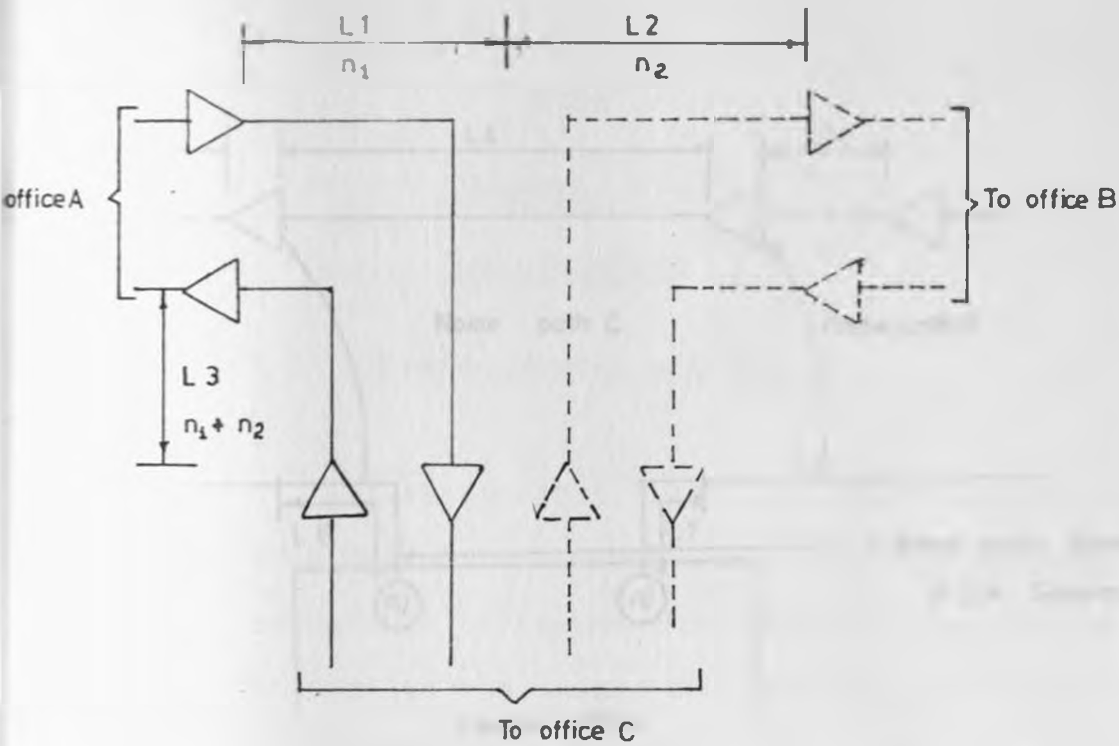


Fig 5·3 Junction of one-cable systems

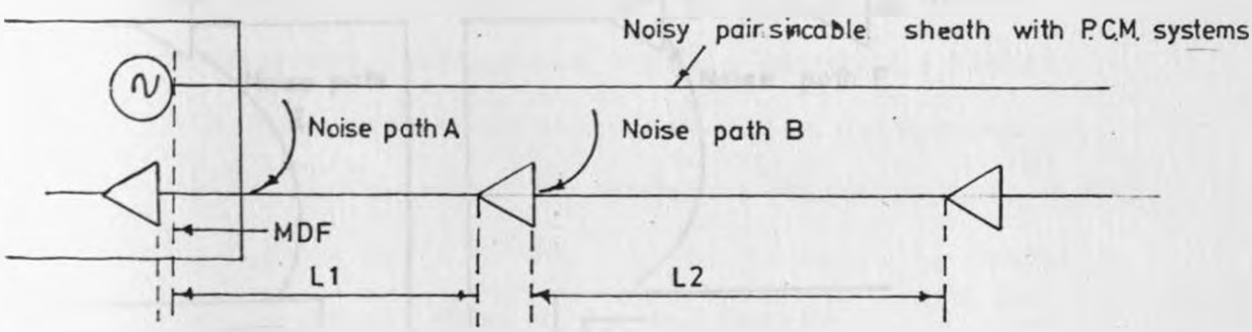


Fig 5·4 Reaper sections near a central office with PCM Systems entering the office.

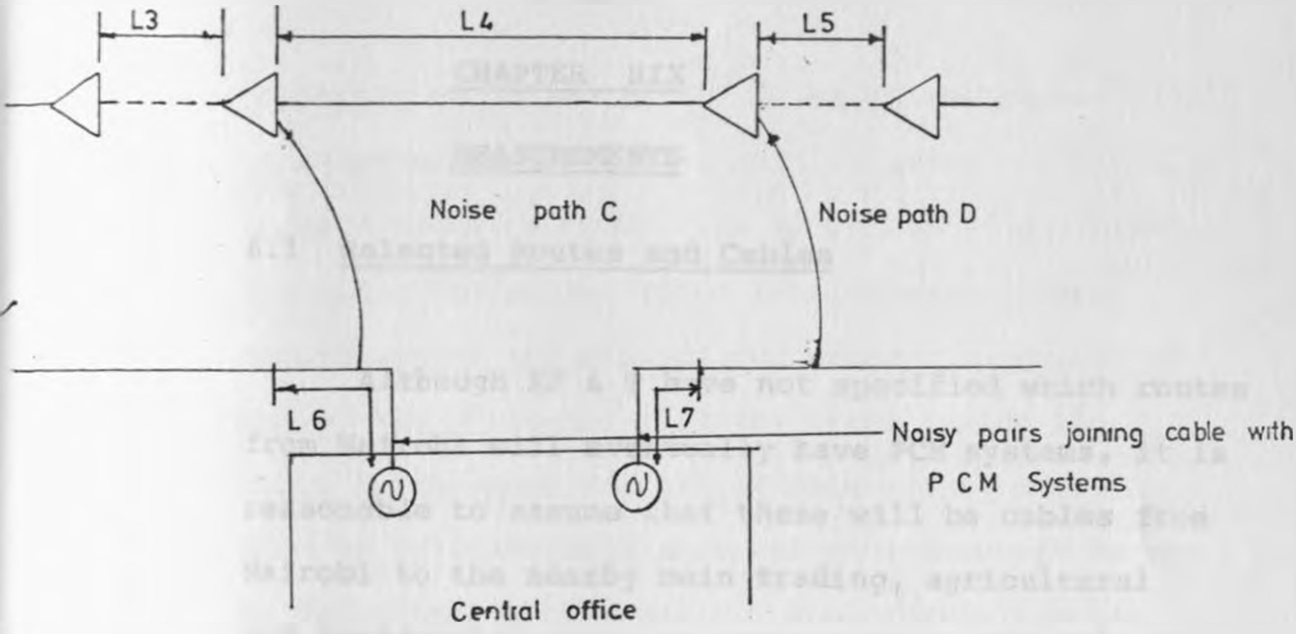


Fig 5.5 Repeater sections near an office with P C M Systems not entering the office

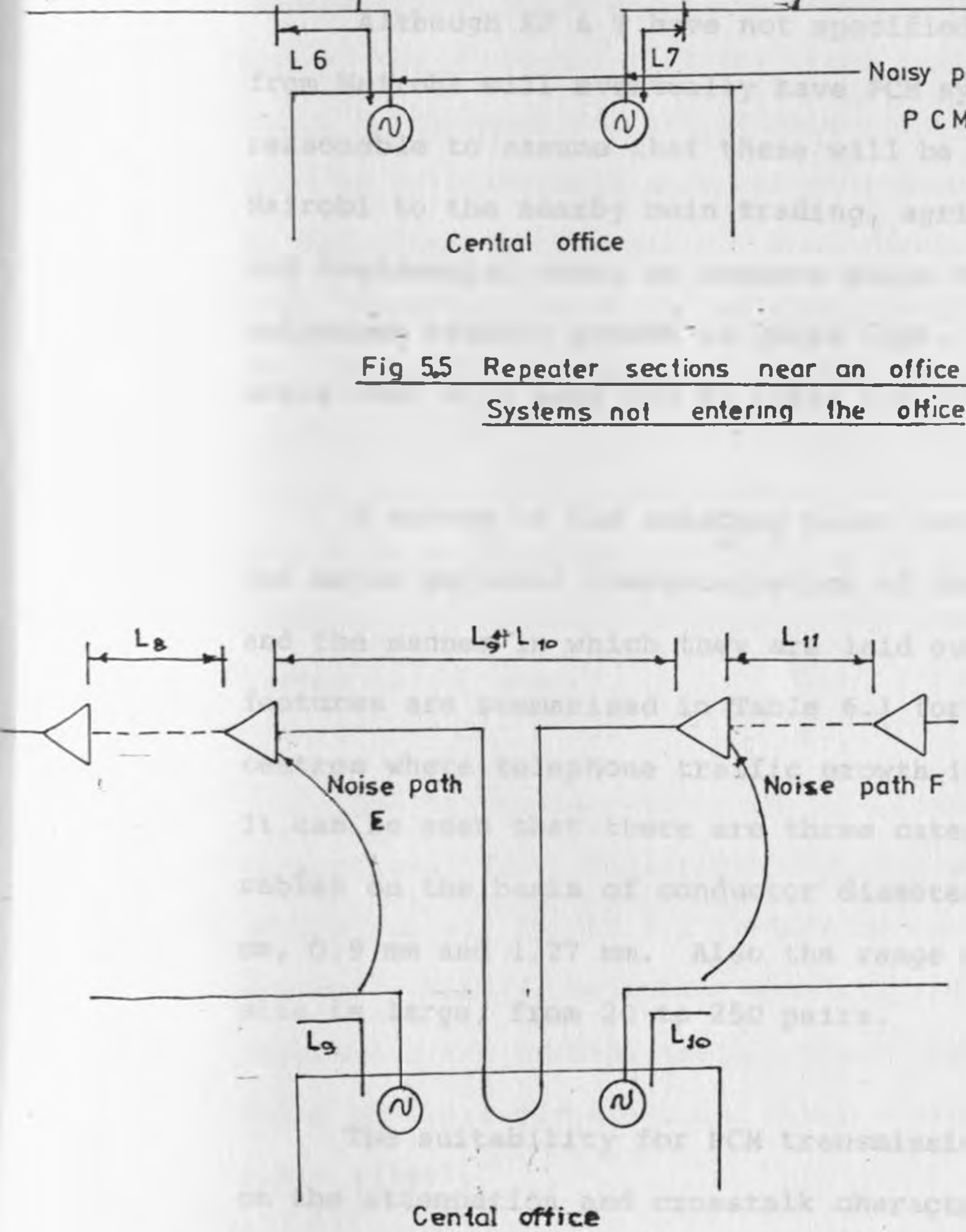


Fig 5.6 Repeater sections near an office with PCM Pairs cross-connected at office main distribution frame

CHAPTER SIX

MEASUREMENTS

6.1 Selected Routes and Cables

Although KP & T have not specified which routes from Nairobi will eventually have PCM systems, it is reasonable to assume that these will be cables from Nairobi to the nearby main trading, agricultural and residential areas or centres where the rate of telephone traffic growth is quite high. It is these areas that will need PCM to cater for the high traffic.

A survey of the existing plant records shows the major physical characteristics of the cables used and the manner in which they are laid out. These features are summarised in Table 6.1 for seven centres where telephone traffic growth is highest. It can be seen that there are three categories of cables on the basis of conductor diameter i.e. 0.63 mm, 0.9 mm and 1.27 mm. Also the range of cable size is large, from 20 to 250 pairs.

The suitability for PCM transmissions depends on the attenuation and crosstalk characteristics of the cable. In turn, attenuation depends on cable diameter while crosstalk noise depends on the relative positions of GO and RETURN pairs in the cable. It is therefore necessary to test at least one cable in each diameter group. In addition, it has been found for the cables under consideration, that regardless of the size of the cable i.e. number of pairs, the arrangement of the wires inside the cable is the same for all of them. They are all made up of star-quads arranged in concentric layers as described in Appendix E. The crosstalk noise will therefore depend on the combination of the layers for the GO and RETURN pairs and not on the overall number of pairs in the cable. Thus results obtained on one cable of a given diameter, can be used to predict the performance of another cable with similar wire diameter as long as the layout of the layers is the same.

Due to considerations on telephone traffic disruptions, time limitations and easy accessibility to the cables, the KP & T decided on cables : Nairobi - Kikuyu, Nairobi - Embakasi and Nairobi - Kiambu for the initial tests. These represent the three possible diameters and cover a large range of cable sizes.

Route	Length(km)	Longest Section (km)	Number of Loading Points	Number of Pairs in Cable	Conductor Diameter (mm)
Nairobi - Embakasi	15.68	1.84	9	250	0.63
Nairobi - Kikuyu	22.2	1.84	12	54	0.9
Nairobi - Muthaiga	6.77	1.84	4	54	0.9
Nairobi - Langata	14.5	1.95	8	44	0.9
Nairobi - Eastleigh	6.14	1.83	3	100	0.63
Nairobi - Athiriver	28.85	1.82	16	54	0.63
Nairobi - Kiambu	15.72	1.83	9	20	1.27

Table 6.1 Features of cable routes considered in the project

6.2 Measurement Procedure

For each of the above cable, measurements were made, wherever possible on the longest section along the route, since it is this section with highest cable attenuation, and would therefore govern the maximum number of PCM systems in the cable. In cases where the longest section could not be used (say when one or both of the manholes for loading points are in the middle of the street); then the next longest section is used.

Before making any measurements, the cable section has to deloaded by disconnecting all pairs under test from the loading.coils on either end of the section. Then using a test tone, the individual pairs are identified and marked on each end of the section. A test telephone connected to one pair in the cable is used for communication between people at each end of the section. The three measurements, mainly cable attenuation, NEXT and FEXT noise were carried out as follows:

6.2.1 Cable Attenuation:

The output marked HDB3 OUT of one of the two PCM test sets is connected to the pair under test and at the other end of the section, the same pair is connected to the PCM INPUT of the second test set, as shown in fig. 6.1. The cable attenuation for this pair is obtained by turning a knob marked CABLE ATT. on the latter test set, clockwise from 4.5 dB until a light marked L1 lights. The attenuation in dB is read off from the dial of the CABLE ATT. knob.

The cable attenuation is measured on about 10 randomly chosen pairs in the cable and the mean value Lo dB is computed.

6.2.2 Near End Crosstalk

As it has already been pointed out, crosstalk noise depends on the relative positions of the "disturbing" and "disturbed" pairs within the cable. The cables under investigation are all of the star-quad multi-layer type in which five different modes of cable filling can be distinguished as was explained in section 5.3.2. It is therefore necessary to obtain values of NEXT noise for each of the modes in order to determine which one is most suitable under given circumstances.

Mode A: In this mode, the "disturbing" and "disturbed" pairs must lie in separate and non-adjacent layers. This means that the cable must have at least 3 layers or according to Appendix E, a minimum of 120 pairs. NEXT measurements for this mode were therefore carried out on the Nairobi-Embakasi cable and on a section between Nairobi and Baldwin (on the Nairobi-Kikuyu cable); which had 150 pairs.

Mode B: In mode B, the "disturbing" and "disturbed" pairs were chosen from any two adjacent layers. Measurements in this mode were carried out on all chosen cables since they all had at least 2 layers.

Mode C: Here the two pairs lie in the same layer but in alternating quads and readings in this mode were taken on all cables.

Mode D: In mode D, the two pairs are in the same layer but in quads separated by one quad. Again readings were taken on all cables.

Mode E: In this mode pairs are chosen randomly without any fixed pattern except that the two pairs must be in different quads.

Fig. 6.2 shows the circuit used for the measurement of NEXT noise for any of the five modes. The HDB3 OUTPUT terminal of the PCM test set is connected to the "disturbing" pair at one end of the section. The other end of the pair is terminated with a 120-Ω resistor. This termination is necessary to mutch the line to the test set. (The characteristic impedance of a pair of conductors in the cable is 120Ω). A selective level measuring set capable of measuring small signals at frequencies of over 1MHz (say a valve voltmeter) is connected to the near end of the "disturbed" pair i.e. the end nearest to the PCM test-set. The other end of the "disturbed" pair is also terminated in 120Ω. When the "disturbing" pair is energized with a PCM signal the NEXT noise induced in the "disturbed" pairs is given by the voltmeter in dBm, which is then expressed in dB relative to the disturbing signal, whose magnitude is known from the test set.

Ideally, the procedure would be to choose a "disturbing" pair from one layer and meausre the induced NEXT noise in ALL other pairs in the same or different layer depending on the mode. Then another "disturbing" pair would be chosen and a similar process repeated and so on. This would lead to an excessively large number of readings and would be too time consuming. However, inorder to ensure statistical stability, the couplings measured for any mode are made as many as possible (in many cases more than 50) and are made between pairs chosen as randomly as possible from layers of interest. From these readings, the mean near end crosstalk M_N and the standard deviation σ_N, are calculated for each mode.

6.2.3. Far-End Crosstalk

As it has already been pointed out in section 5.4, knowledge of far-end crosstalk noise is necessary only in a two-cable operation, and since KP & T will operate a one-cable system initially, FEXT measurements were made to provide a comparison between one-cable and two-cable operation.

The procedure is similar to that used for NEXT measurements except the voltmeter has to be connected to the far end of the "disturbed" pair, whose near end is terminated in a 120Ω-resistor. The circuit diagram used is shown in fig. 6.3.

Again to ensure statistical stability, the couplings measured have to be as many as possible and have to be made between pairs chosen as randomly as possible from the relevant layers. From the readings, the mean far-end crosstalk M_F and the standard deviation σ_F, for each mode, are calculated.

6.3 <u>Results</u>:

The mean section loss L_o dB was computed using the formula

$$L_o = \frac{\sum_{i=1}^{n} L_i}{n}$$

where L_i represents the cable attenuation for one pair and n is the number of pairs measured.

Similarly, for each of the five modes, the mean NEXT noise, M_N, and mean FEXT noise, M_F, were computed using the formulae:

$$M_N = \frac{\sum_{i=1}^{n} X_i}{n} \quad \text{and} \quad M_F = \frac{\sum_{i=1}^{n} X_i}{n}$$

where X_i represents the individual couplings in dB and n, the number of couplings measured.

The standard deviation σ_N and σ_F for the NEXT and FEXT noises respectively were obtained from:

$$\sigma_N = \sqrt{\frac{\sum_{i=1}^{n} (X_i - M_N)^2}{n}} \quad \text{and} \quad \sigma_F = \sqrt{\frac{\sum_{i=1}^{n} (X_i - M_F)^2}{n}}$$

Using the computed values of L_o, M_N, M_F, σ_N and σ_F for each mode and the curves in fig. 5.1, the maximum number of systems n for a one-cable and a two-cable operation, can be found. It should be noted that in finding n, an appropriate error margin M_{ε} dB and a correction factor $\Delta(L_o, RC)$, dB have to be taken into account.

For the number of systems limited by NEXT noise, the correction factor $\Delta(L_o, RC)$ dB for section attenuation other than 20 dB, is obtained from the curve in fig. 5.2. This correction does not apply to the FEXT case. A recommended error margin of 8 dB (6) is used in both cases. In the NEXT case, the sum of the error margin and the correction factor is subtracted from M_N to obtain a modified value of mean NEXT noise which is used to obtain n. The modified value of mean FEXT that is used in fig. 5.1 to determine n, is obtained by subtracting the error margin and a factor of 0.3 (explained in section 5.4), from the computed M_F.

The following tables show results summarised for the three cables on which measurements were made.

(i) Nairobi-Kikuyu

Number of pairs: 54
Conductor diameter: 0.9 mm
Section length: 1.84 km.
Mean section loss L_o: 24 dB.

Mode	Mean NEXT M_N(dB)	Standard Deviation σ_N(dB)	Max. No. of Systems n
A	90.6	8	300
B	71.2	7	28
C	68.1	7	8
D	65.2	7	3
E	68.3	11	0

Table 6.2: Maximum number of systems limited by NEXT for a 0.9 mm cable.

Mode	Mean FEXT M_F(dB)	Standard Deviation σ_F(dB)	Max. No. of Systems n
A	73.3	12	300
B	68.4	11	250
C	63.2	10	200
D	58.3	10	30
E	63.4	11	50

Table 6.3: Maximum number of systems limited by FEXT for a 0.9 mm cable. (Nairobi-Kikuyu).

(ii) Nairobi - Embakasi

Number of pairs: 250
Conductor Diameter: 0.63 mm
Section length: 1.84 km.
Mean section loss L_o: 35 dB.

Mode	Mean NEXT M_N(dB)	Standard Deviation σ_N(dB)	Max. No. of Systems n
A	83.7	7	42
B	75.6	6	5
C	73.7	6	2
D	68.5	7	0
E	70.8	10	0

Table 6.4: Maximum number of systems limited by NEXT for a 0.63 mm cable. (Nairobi-Embakasi)

Mode	Mean FEXT M_F(dB)	Standard Deviation σ_F(dB)	Max. No. of systems n
A	72.3	12	250
B	66.2	11	130
C	64.1	11	64
D	53.4	10	7
E	58.3	11	14

Table 6.5: Maximum number of systems limited by FEXT for a 0.63 mm cable. (Nairobi-Embakasi)

(iii) Nairobi - Kiambu

Number of pairs: 20

Conductor diameter: 1.27 mm

Section length: 1.84 km

Mean section loss L_o: 15 dB.

Mode	Mean NEXT M_N(dB)	Standard Deviation σ_N(dB)	Max. No. of systems n
B	70.1	8	170
C	68.0	7	240
D	63.2	7	47
E	65.1	10	3

Table 6.6: Maximum number of systems limited by NEXT for a 1.27 mm cable. (Nairobi-Kiambu)

Mode	Mean FEXT M_F(dB)	Standard Deviation σ_F(dB)	Max. No. of Systems n
B	72.3	12	250
C	66.2	11	130
D	62.4	11	40
E	65.3	11	90

Table 6.7: Maximum number of systems limited by FEXT for a 1.27 mm cable. (Nairobi-Kiambu)

CHAPTER SEVEN

DISCUSSION

7.1 Comments on Measurements.

It is clear from the computed mean NEXT, M_N, and mean FEXT, M_F, that for all cable diameters, mode A has the best crosstalk performance followed by mode B,C, E and D in that order (see figs. 6.4 and 6.5). This is to be expected since the degree of separation between the "disturbed and "disturbing" pairs is highest in mode A and decreases progressively in modes B, C and D. In mode E, the separation varies randomly, since pairs are chosen without any restriction, provided they are not in the same quad.

However, the standard deviation for NEXT noise tends to be highest in mode E, followed by modes A, D, C and B is shown in Fig. 6.6 while for FEXT noise it is highest in A and E and falls to lower values in B, C and D. For a given mean crosstalk, the maximum number of systems in a cable decreases as the standard deviation (or spread in the crosstalk values) increases, (see fig. 5.1). It turns out that, though mode A has a relatively high standard deviation, its excellent crosstalk values enables it to have the highest number of systems followed by modes B, C, D and E in that order for NEXT noise. The high spread in E, limits the number of systems even though its crosstalk is fairly good (better than D's).

Though A has the best performance, it has a serious disadvantage in that a maximum of only one half of the pairs in the cable can be used for PCM transmission and the cable has to be large (at least 120 pairs). In D also, the maximum cable fill up is 50%, a fact when coupled with its poor crosstalk performance makes it the poorest choice of the 5 modes. Modes B, and C have all a maximum fill up of 100% making them fair alternatives compared to A (as shown in Tables 6.8 to 6.13).

The section loss L_o, drastically reduces the possible number of systems (in a one-cable operation) as it increases. This can be seen for the 0.63 mm cable which has the highest loss per km. This cable has the lowest number of systems for any mode with no PCM transmission possible in modes D and E. The only way to use this cable in these modes is to reduce the section losses by reducing the section lengths. This will however mean digging new manholes where regenerators will be positioned and an increase in the total number of regenerators, an alternative which may prove too expensive. Otherwise the only modes that can be used for this cable are A, B and C. Also for the 0.9 mm cable, mode E cannot be used with a section loss of 24 dB. Because of its low attenuation loss, the 1.27 mm cable, may be used in any mode using the existing manholes for the regenerators. It should be noted that in the two-cable operation (where FEXT controls), the section loss does not matter.

The choice of what mode to use will obviously depend on the number of systems required, with mode A giving the highest followed by B, C, E and D respectively. If however, the number required can be provided by E, this mode should be used since no strict arrangement of pairs to be followed. It is also the only mode that can be used if the arrangement of pairs in a cable is not uniform over a section.

It is evident from Tables 6.2 to 6.7, that though FEXT has slightly lower mean values and larger standard deviations than NEXT, for all modes, it permits transmission of more PCM systems. This is of course at the expense of laying an extra cable for each route, a project that KP & T is not undertaking now. The FEXT measurements, as mentioned before, are presented to provide a basis for comparison between one-cable and two-cable operation. This will be useful should demand for more systems, than those provided by one-cable operation, arise in future.

The following 6 tables show the improvement in the number of telephone channels along each route considered in this project, if one-cable or two-cable operation is used under any given mode. In the existing systems, the maximum number of telephone channels equals the number of pairs in the cable less two pairs which are used for supervisory and fault locating purposes. A two-wire circuit is used for each channel. With the proposed type of PCM transmission, each PCM system has 32 channels, 2 of which are used for frame alignment and signalling, leaving 30 telephone channels per system (see section 4.1).

(i) Nairobi-Kikuyu (54 pairs, 0.9 mm), Nairobi-Muthaiga (54 pairs, 0.9 mm)

Mode	Existing No. of Channels	Max. No. of Channels with one-cable operation	Max. No. of Channels with 2.cable operation
A	52	N/A	N/A
B	52	810	1620
C	52	240	1620
D	52	90	900
E	52	NONE	1500

Table 6.8: Number of telephone channels for a 54-pair, 0.9 mm cable with one-cable and two-cable operation.

(ii) Nairobi-Langata (44 pairs, 0.9 mm)

Mode	Existing No. of Channels	Max. No. of Channels with one-cable Operation	Max. No. of Channels with two-cable Operation
A	42	N/A	N/A
B	42	660	1320
C	42	240	1320
D	42	90	900
E	42	NONE	1320

Table 6.9: Number of telephone channels for a 44-pair, 0.9 mm cable with one-cable and two-cable operation.

(iii) Nairobi-Embakasi (250 pairs, 0.63 mm)

Mode	Existing No. Channels	Max. No. of Channels with one-cable operation	Max. No. of Channels with two-cable operation
A	248	1260	2250
B	248	150	3900
C	248	60	1920
D	248	NONE	210
E	248	NONE	420

Table 6.10: Number of telephone channels for a 250-pair, 0.63 mm cable with one-cable and two-cable operation.

(iv) Nairobi-Eastleigh (100 pairs, 0.63 mm)

Mode	Existing No. of Channels	Max. No. of Channels with one-cable operation	Max. No. of Channels with two-cable operation
A	98	N/A	N/A
B	98	150	3000
C	98	60	1920
D	98	NONE	210
E	98	NONE	420

Table 6.11: Number of telephone channels for a 100-pair, 0.63 mm cable with one-cable and two-cable operation.

(v) Nairobi-Athi River (54 pairs, 0.63 mm)

Mode	Existing No. of Channels	Max. No. of Channels with one-cable operation	Max. No. of Channels with two-cable operation
A	52	N/A	N/A
B	52	810	1620
C	52	150	1620
D	52	NONE	210
E	52	NONE	420

Table 6.2L: Number of telephone channels for a 54-pair, 0.63 mm cable with one-cable and two-cable operation.

(vi) Nairobi-Kiambu (20 pairs, 1.27 mm)

Mode	Existing No. of Channels	Max. No. of Channels with one-cable operation	Max. No. of Channels with two-cable operation
A	18	N/A	N/A
B	18	300	600
C	18	300	600
D	18	300	600
E	18	90	600

Table 6.13: Number of telephone channels for a 20-pair, 1.27 mm cable.

Considering one-cable operation only, from Tables 6,8 and 6.9 it is clear that the only modes that can be used for the Nairobi-Kikuyu, Nairobi-Muthaiga and Nairobi-Langata cables, with any significant improvement in telephone channels are B and C. With B, there is an improvement factor of about 16 while for C it is 5. E cannot be used unless the repeater sections are shortened. The improvement factor with D (about 2) is too small to justify the increase in costs caused by terminal equipment.

For the Nairobi-Embakasi cable, only the A mode may be used with an increase to about five times the number of existing channels. Because of the high attenuation of this cable, other modes are unsuitable, unless the repeater sections are made shorter. The same arguement holds for other 0.63 mm cables, where mode B is the only one suitable for the Nairobi-Eastleigh cable (improvement factor ≃ 1.5) and modes B and C for the Nairobi-Athiriver cable.

Though the 1.27 mm cable has a very good improvement factor (over 15 for modes B, C and D and 5 for E) this cable is heavy and hence has very few pairs. But is is quite suitable for PCM transmission for all modes.

It should be noted that although all the cables considered terminate at the MDF on either end, the errors that are due to office noise are minimised because all terminal sections are one half (or less) the length of the intermediate sections. Thus because the terminal section loss is equal to one half (or less) that of the intermediate section, the error rate objective will automatically be satisfied as explained in section 5.6. In other words, it is the longest section along the route, which controls the maximum number of PCM systems.

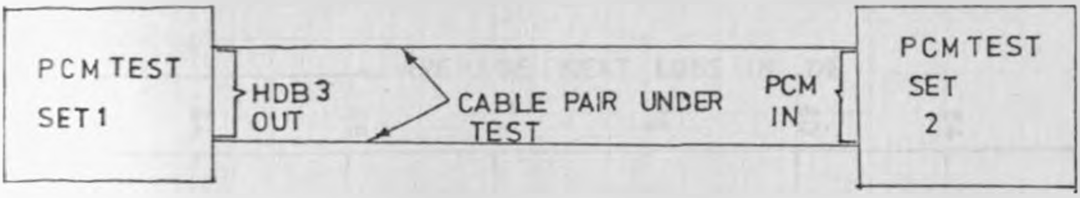


Fig 6·1 Measurement of cable attenuation

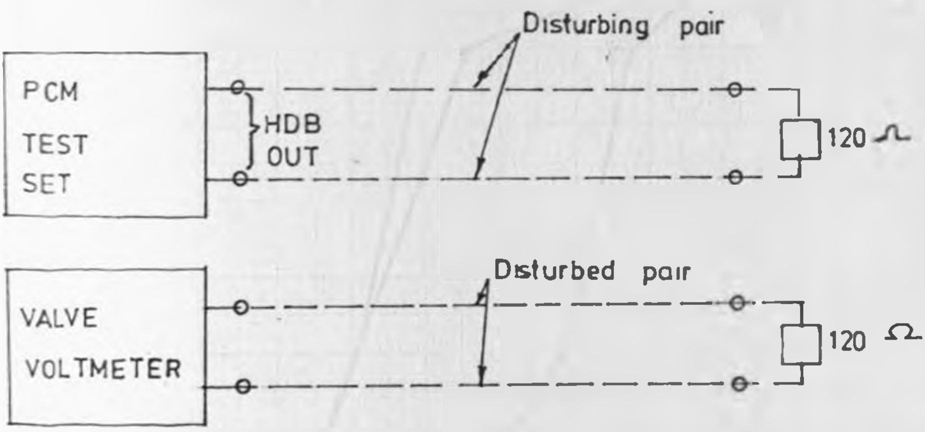


Fig 6·2 Measurement of Near end crosstalk

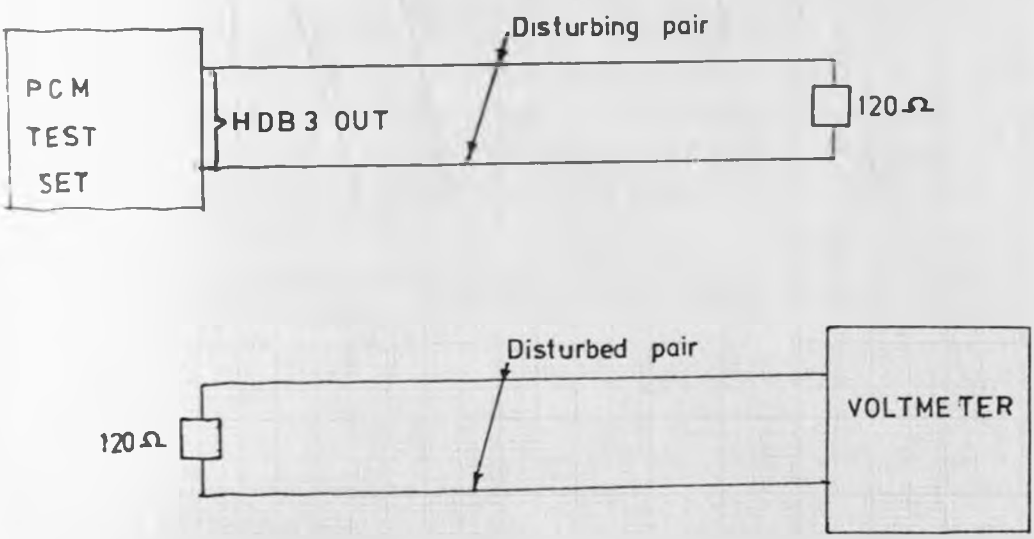
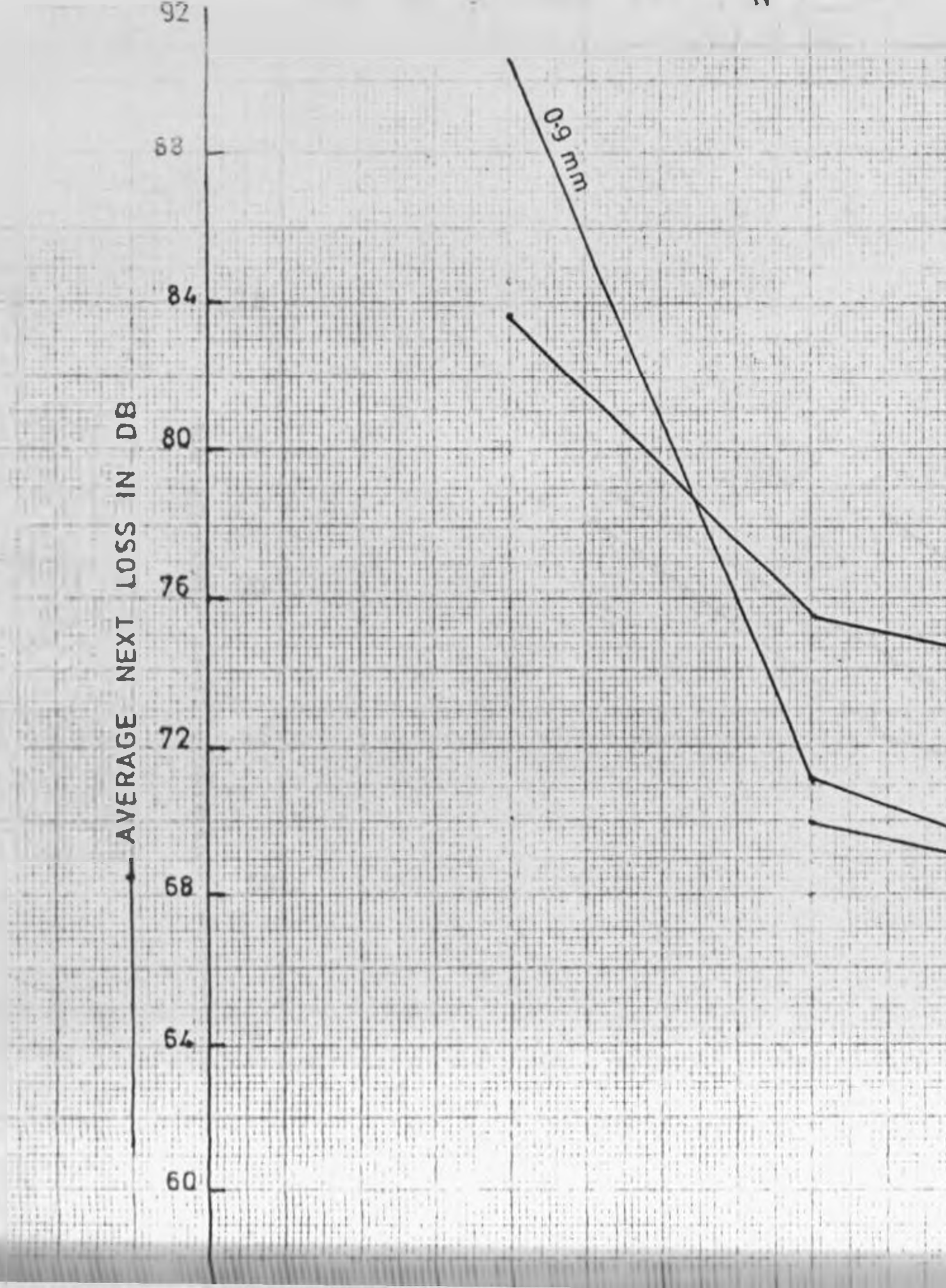


Fig 6·3 Measurement of far end crosstalk

FIG 6·4 AVERAGE NEXT (M_N) VERSUS



LAYER SEPARATION FOR VARIOUS CABLE DIAMETERS

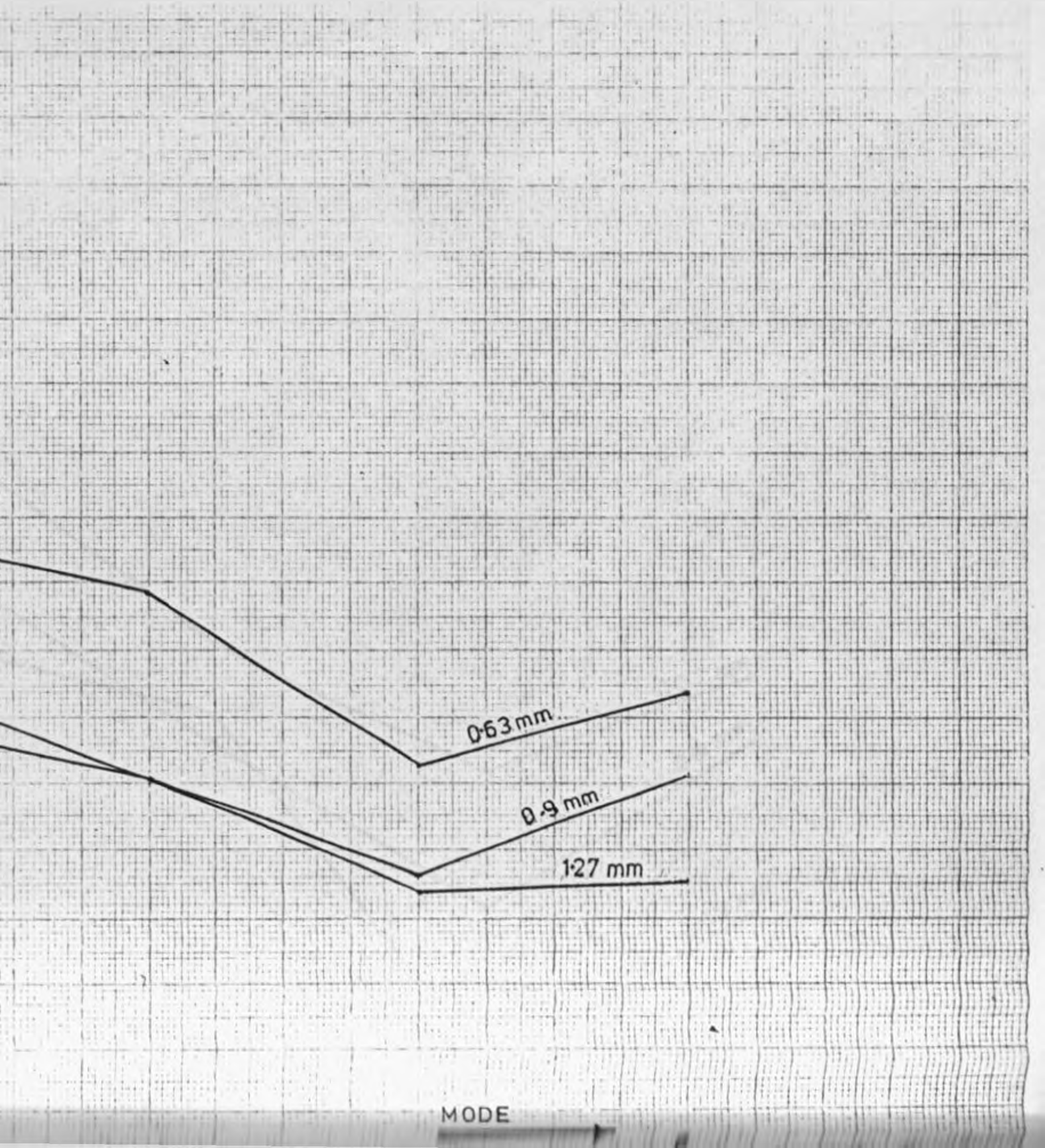


FIG 6·5 AVERAGE FEXT (M_F) VERSUS LAYER SEPARATION FOR VARIOUS CABLE DIAMETERS

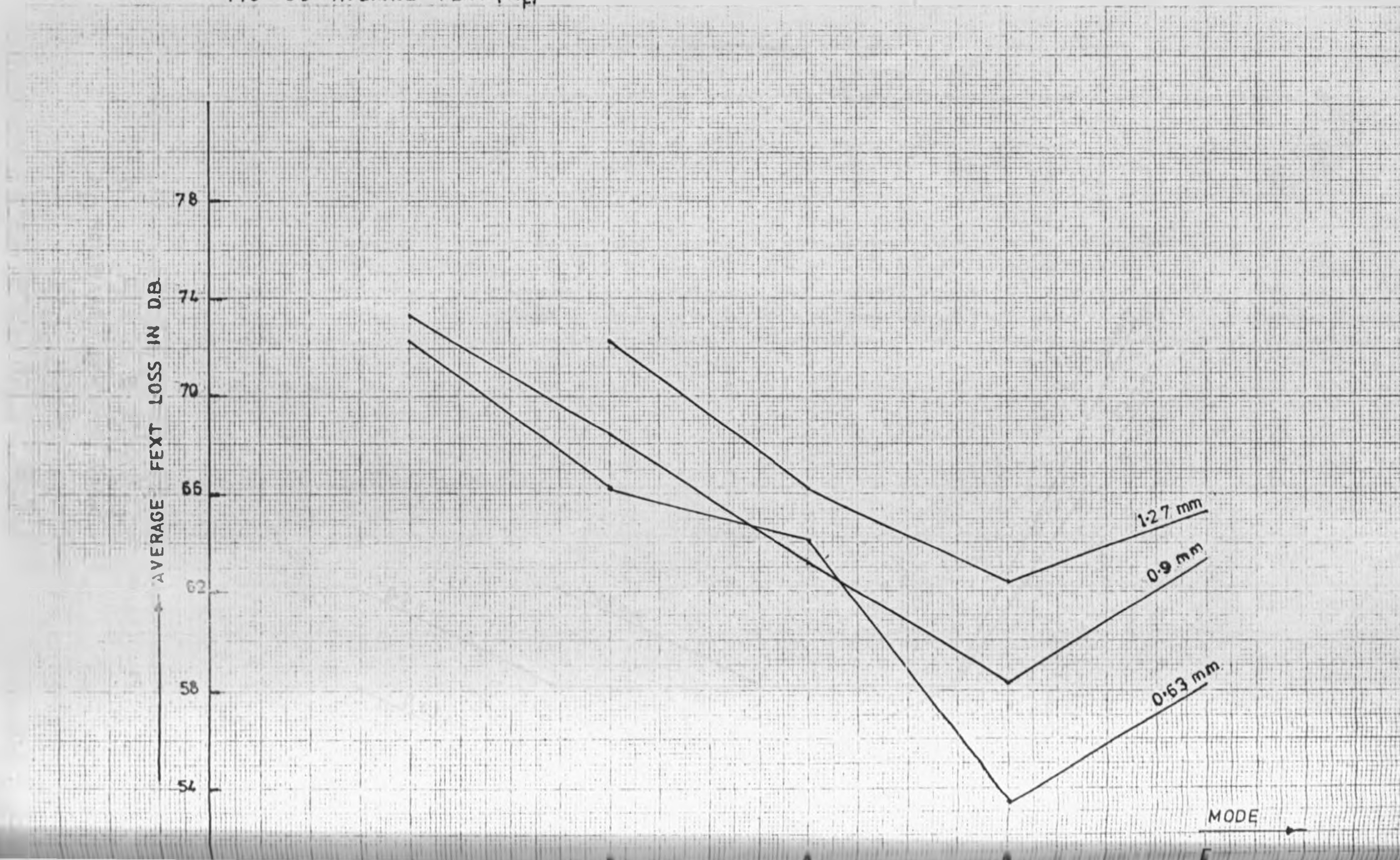
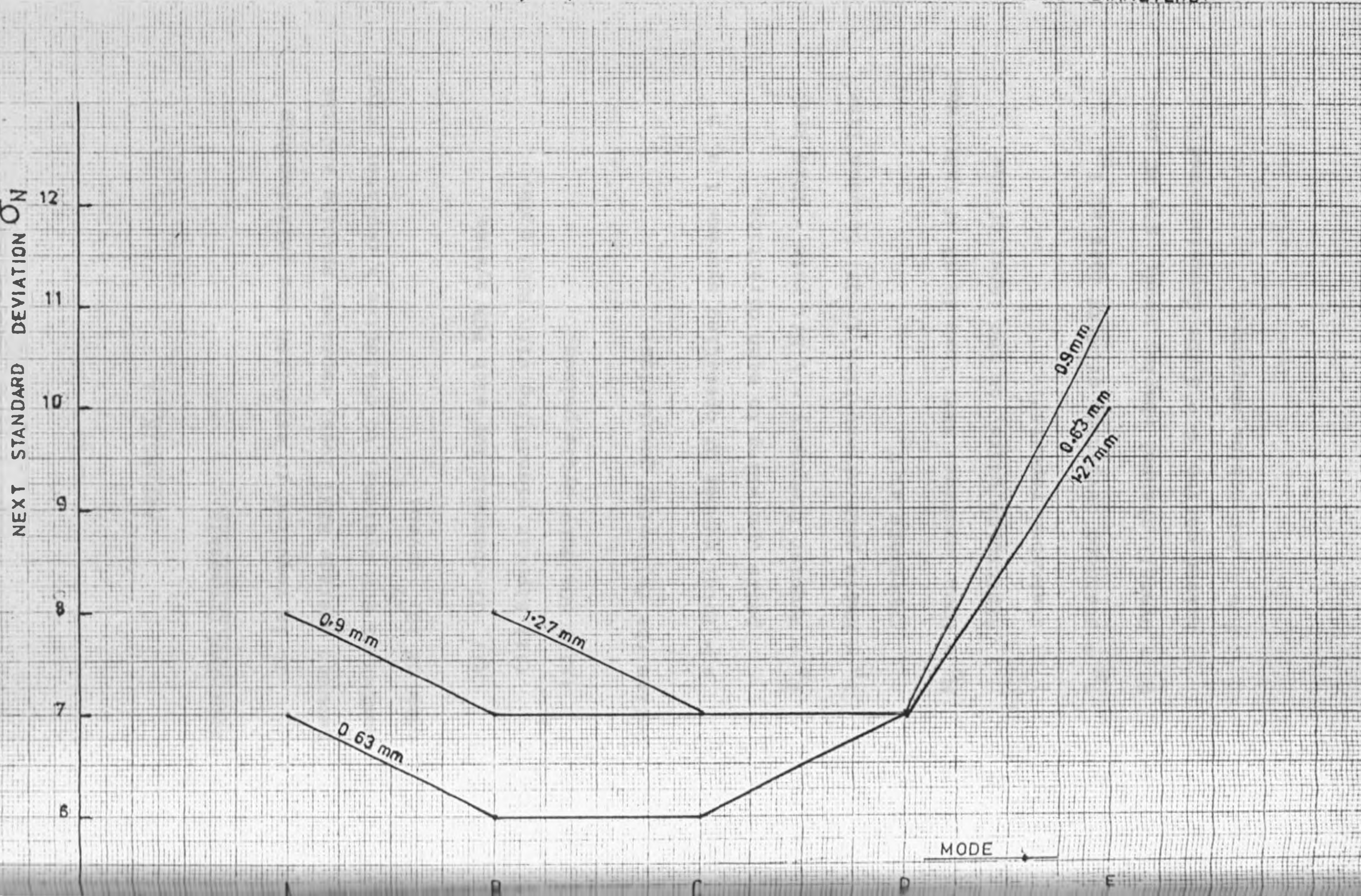


FIG 6·6 NEXT STANDARD DEVIATION, σ_N, VERSUS LAYER SEPERATION FOR VARIOUS DIAMETERS.



APPENDIX A

Near-End Crosstalk

It is required to establish the relationship between average NEXT loss, repeater spacing and the number of PCM systems, to guarantee a repeater error rate of 10^{-7} given:

(i) 32 channels for each PCM system.

(ii) Bipolar signalling with HDB3 code.

(iii) 50% duty cycle pulses.

Define the following quantities:

n = Total number of operating systems i.e. n pairs operating in each direction.

f_o = Nyquist frequency = ½ bit rate = 1024 kHz.

L_o = Cable pair attenuation in dB, between line repeaters at f_o.

M_N = Mean (average) NEXT loss in dB of individual disturbing pairs, at the input of the disturbed repeater.

σ_N = Standard deviation of NEXT loss at f_o.

S = Overall channel shaping used; S = RC for raised cosine, S = G for Gaussian, and S = C for cosine shaping.

Consider fig. A.1, which is the NEXT model to be used for analysis. Let,

C(f) = Cable response to a rectangular pulse.

H(f) = Equalizer gain to compensate for cable loss and provide required channel shaping.

R(f) = Amplitude spectrum of the equalized pulse.

G(f) = Amplitude spectrum of a 50% duty-cycle rectangular pulse, with base-to-peak value of V_t volts.

P(f) = Power spectrum for the disturbing bipolar signal with equally likely 1's and 0's in the binary stream

From Fig. A.1, the equalized pulse is given by:

$$|R(f)|^2 = |G(f)|^2 \; |C(f)|^2 \; |H(f)|^2$$

Thus the equalizer response is:

$$|H(f)|^2 = \frac{|R(f)|^2}{|C(f)|^2 \; |G(f)|^2} = \frac{R^2}{G^2 \; C^2} \text{ ---- (A.1)}$$

If $X_N(f)$ = power transfer function through the NEXT path, then the interference power Q dB, at the output of the equalizer for the frequency range $0\text{-}2f_o$ (since the disturbing bipolar pulse train has a continuous spectrum in this frequency range) is given by

$$10^{0.1Q} = \int_0^{2f_o} P(f) X_N^2 H^2 df \quad \text{watts} \text{ -------- } (A.2)$$

Now if, the mean NEXT loss at f_o is M_N dB, then the NEXT transfer function $X_N(f)$ is given by (6).

$$|X_N(f)|^2 = 10^{-M_N/10} \, (f/f_o)^{3/2} \text{ ---------- } (A.3)$$

Also, the power spectrum P(f), of the disturbing bipolar signal is given by (8).

$$P(f) = \frac{2f_o}{Z_o} \quad |G(f)|^2 \, (1 - \mathrm{Cos}^{\pi f}/f_o)$$

where, Z_o = characteristic impedance of the cable.

The cable response C(f) to a pulse response is given by (6)

$$|C(f)|^2 = \exp\{-\alpha_0 \sqrt{f/f_o}\}, \text{ ------------- } (A.5)$$

where α_0 = cable attenuation in nepers at f_o.

Substituting in (A.2) for P(f), $X_N(f)$ and H(f), the interference power at the equalizer output due to one interferring system is

$$10.^{01Q} = \int_o^{2f_o} \frac{2f_o}{Z_o} \, |G(f)|^2 (1 - \mathrm{Cos}^{\pi f}/f_o)$$

$$\times \quad \frac{R^2 \, 10^{-M_N/10} (f/f_o)^{3/2}}{G^2 \exp\{-\alpha_0 \sqrt{f/f_o}\}} \; df$$

Denoting this power relative to 1 watt as $Q_N(1)$, we have ,

$$Q_N(1) = -m_N + 10 \log \int_0^{2f_o} R^2 \frac{4f_o}{Z_o} (f/f_o)^{3/2} \cdot$$

$$\exp\{\alpha_0 \sqrt{f/f_o}\} \operatorname{Sin}^2 2B \, df \text{ dBW.} \text{ ------(A.6)}$$

where $B = \dfrac{\pi f}{4f_o}$

It is seen from equation (A.6), that the NEXT interference power from one system depends on the cable attenuation and the channel shaping used. It can be simplified by noting that for any type of channel shaping, the specturm of the equalized pulse is given by

$$|R(f)|^2 = \frac{V_r^2}{Z_o} \quad |A(f)|^2 \text{ where}$$

V_r = base-to-peak value of the equalized signal and

A(f) is some function of frequency.

If $D_N(L_o,S)$ dBW, is the normalized NEXT interference (ie, $Q_N(1) = D_N(L_o,S)$, when $M_N = 0$ and $\dfrac{V_r^2}{Z_o} = 1$), for cable attenuation L_o and channel shaping S, then the NEXT interference power from one system may be written as

$$Q_N(1) = D_N(L_o,S) + 10 \log \frac{V_r^2}{Z_o} - M_N, \text{dBW} \text{ -------(A.7)}$$

When there are n systems in operation in one cable, it can be shown that the total NEXT power $Q_N(n)$ dBW at any repeater is(6).

$$Q_N(n) = -M_N + D_N(L_o,S) + 10\log\frac{V_r^2}{Z_o} + I(\sigma_N,n)$$

$$+\ 2.33\sigma_\phi(\sigma_N,n) \quad \text{-------------} \quad (A.8)$$

where $I(\sigma_N,n)$ is the increase in NEXT power over that due to a single source and is given by

$$I(\sigma_N,n) = 5\Big[\log(n^3\exp\{0.05\sigma_N^2\}) - \log(\exp\{0.053\sigma_N^2\}$$

$$+\ n-1)\Big] \quad \text{--------------------} \quad (A.9)$$

and $$\sigma_\phi(\sigma_N,n) = 6.593\Big[\log(\exp\{0.053\sigma_N^2\}$$

$$+\ n-1) - \log n\Big]^{\frac{1}{2}} \quad \text{--------------------} \quad (A.10)$$

It has been assumed in equation (A.8) that the interference power from the n systems has a Gaussian distribution. The NEXT power level at 99% of the repeaters will be less or equal $Q_N(n)$ in equation (A.8).

Since the peak equalized signal power is V_r^2/Z_o, it follows that the peak signal power to average NEXT interference power (S/N ratio) is:

$$S/N = 10 \log \frac{V_r^2}{Z_o} - \left[-M_N + D_N(L_o,S) + 10 \log \frac{V_r^2}{Z_o} \right.$$

$$\left. + M_\varepsilon + I(\sigma_N,n) + 2.33\sigma_\phi(\sigma_N,n) \right] \quad \text{------} \quad (A.11)$$

Where M_ε is the error margin in dB, allocated for other noise sources such as intersymbol interference or misequalization, thermal noise, timing jitter and incomplete knowledge of crosstalk data.

Assuming the total interference to be Gaussian, a S/N ratio of 20.8 dB results in an average error rate of 10^{-7} for the detection of 3 levels (as in HDB3 code) (6). Thus to guarantee an error rate of 10^{-7},

$$M_N - D_N - I - 2.33\sigma_\phi \geq 20.8 + M_\varepsilon$$

or subtracting cable attenuation L_o from both sides,

$$M_N - L_o \geq 20.8 + \left[D_N - L_o \right] + \left[I + 2.33\sigma_\phi \right] + M_\varepsilon \quad \text{--------} (A.12)$$

It is convenient to assign nominal values to L_o and S in order to determine $D_N(L_o,S)$. Plots of M_N versus n can then be made for different values of σ_N. A correction factor $\Delta_N(L_o,S)$ is introduced if different values of L_o and S are used.

For example: $D_N(L_o, RC)$ given by;

$$D_N(L_o,RC) = 10 \log \int_0^{2f_o} \frac{1}{f_o} \left(\frac{f}{f_o}\right)^{3/2} .$$

$$\exp\{\alpha_0 \sqrt{f/f_o}\} \sin^2 2B \cos^2 B \, df \text{ ---------(A.13)}$$

has been calculated by numerical integration (6) for $L_o = 43$ dB and S = raised cosine to give $D_N(43,RC)$ = 36.5 dB.

For any other cable attenuation L_o, and channel shaping S, equation (A.12) can then be written as

$$M_N - L_o \geq 20.8 + \left[D_N(43,RC) - 43\right] + \left[I + 2.33\sigma_\phi\right]$$
$$+ \Delta_N(L_o,S) + M_\varepsilon \text{ --------------(A.14)}$$

where the correction $\Delta_N(L_o, S)$ is given by,

$$\Delta_N(L_o,S) = \left[43 - L_o\right] - \left[D_N(43,RC) - D_N(L_o,S)\right]$$
$$\text{------------ (A.15)}$$

Equation A.14 simplifies to:

$$M_N - L_o \geq 14.3 + \left[I(\sigma_N,n) + 2.33\sigma_\phi(\sigma_N,n)\right]$$
$$+ \Delta_N(L_o,S) + M_\varepsilon \text{ -----------(A.16)}$$

It should be noted at this point, that the correction factor $\Delta_N(L_o,S)$ for the cable attenuation L_o, specified by KP & T (5-35dB), is very small, ranging from -1.2dB for L_o = 5 dB to 0.6 dB for L_o = 35 dB for raised cosine shaping. It is equally small, for other shaping.

Effect of HDB3 Coding

As explained in Chapter four and Appendix D, in HDB3 code, four consecutive zeros in the random bipolar code are replaced by a predetermined code word to ease the problem of time recovery in the regenerator. This changes the signal power spectrum P(f). Assuming binary 1's and 0's to be equally likely, and raised cosine shaping, the spectral density of an HDB3 signal relative to that of straight bipolar at f is 1.087 (16, 18). This means an increase of 0.36dB in the total interference power $Q_N(n)$. The effect of HDB3 coding on the possible number of PCM systems in a cable, will therefore be accounted for by simply adding 0.36 dB on the RHS of equation (A.16). The modified equation is then,

$$M_N - L_o \geq 14.66 + \left[I(\sigma_N, n) + 2.33\sigma_\phi(\sigma_N, n) \right] + \Delta_N(L_o,S) + M_\varepsilon \quad \text{--------} \quad (A.17)$$

Using equation (A.17), the minimum values of M_N can be plotted against the possible number of systems n, for any given cable when $\Delta_N(L_o,S)$ and M_ε are both zero. Any desired error margin M_ε and correction factor $\Delta_N(L_o,S)$ (which may be neglected without introducing any significance error), may readily be added to the results later. An error margin M_ε of about 6-12 dB is usually added (6).

Since KP & T have specified that for their junction cables, the repeater-section cable attenuation, L_o, at 1MHz, varies between 5 and 35 dB, it is convenient to have curves of mean NEXT loss (M_N) versus the number of maximum PCM systems drawn for nominal section loss L_o = 20 dB for all possible NEXT loss standard deviations $\sigma_N (5 \leq \sigma_N \leq 14)$. Any other section loss different from 20 dB would be catered for by an additional error curve.

The derivation of the curves is as follows:

From equation (A.15), the correction factor $\Delta(L_o,S)$ for a 20 dB section loss relative to a 43 dB section loss, assuming raised-cosine channel shaping, is given by

$$\Delta_N(20,RC) = \left[43-20\right] - \left[D_N(43,RC) - D_N(20,RC)\right] \qquad \text{(A.18)}$$

But from ref (6), $\Delta_N(20,RC) = -1.2$ dB. Using this value, and $D_N(43,RC) = 36.5$ dB , equation (A.18) is solved to give

$$D_N(20,RC) = 12.3$$

Substituting this value in equation (A.12) and adjusting for the effect of HDB3 coding, the NEXT loss M_N and number of systems n, would be governed by

$$M_N - L_O \geq 20.8 + 0.36 + \left[D_N(20,RC) - 20\right] + \left[I + 2.33\sigma_\phi\right] + M_\varepsilon + \Delta_N(L_O,RC).$$

$$\text{or } M_N - L_O \geq 13.46 + \left[I + 2.33\sigma_\phi\right] + M_\varepsilon + \Delta_N(L_O,RC) \qquad \text{------------------(A.19)}$$

Where $\Delta_N(L_O,RC)$ is the correction to be made if a repeater-section loss departs from the nominal value of 20 dB.

Re-arranging equation (A.19), we get

$$M_N - L_O - M_\varepsilon - \Delta_N(L_O,RC) \geq 13.46 + \left[I + 2.33\sigma_\phi\right] \qquad \text{----- (A.20)}$$

The correction curve for section loss other than, 20 dB is obtained by using equation (A.15), from which the correction for loss L_O relative to the nominal loss of 20 dB, is given by:

$$\Delta_N(L_o,RC)_{20} = (20-L_o) - \left[D_N(20,RC) - D_N(L_o,RC)\right]$$

but $D_N(L_o,RC) = D_N(43,RC) - (43-L_o) + \Delta_N(L_o,RC)_{43}$,

where $\Delta_N(L_o,RC)_{43}$ = correction relative to section loss of 43 dB.

Using the known values of $D_N(43,RC) = 36.5$ and $D_N(20,RC) = 12.3$, we have,

$$\Delta_N(L_o,RC)_{20} = 1.2 + \Delta_N(L_o,RC)_{43} \quad \text{---------} \quad (A.21)$$

Using values of $\Delta_N(L_o,RC)_{43}$ for various section losses L_o, from ref (6), a plot of $\Delta_N(L_o,RC)_{20}$ versus L_o can be made using equation (A.21).

APPENDIX B

Far-End-Crosstalk.

To establish the relationship between average FEXT loss and the number of PCM systems, the analysis is similar to that in Appendix A, except a new FEXT path power transfer function $X_F(f)$ is used.

Let M_F = mean FEXT loss in dB at f_o.

and σ_F = standard deviation of FEXT loss at f_o.

Fig. B.1 is a simple FEXT model (6) to the used for the analysis.

In far-end crosstalk, the disturbing signal passes through the whole cable section; Thus the overall power transfer function $X_F(f)$ through the FEXT path is given by (6).

$$|X_F(f)|^2 = |C(f)|^2 10^{-M_F/10} (f/f_o)^2 \text{----------} \quad (B.1)$$

As in the NEXT case, the interference power Q_F dB at the output of the equalizer is:

$$10^{0.1Q} = \int_0^{2f_o} P(f)|X_F(f)|^2|H(f)|^2 \, df$$

$$\text{or } 10^{0.1Q} = \int_0^{2f_o} P(f)\,|C(f)|^2\, 10^{-M_F/10} ({}^f/f_o)^2$$

$$x\ X \frac{|R(f)|^2}{|G(f)|^2|C(f)|^2}\ df$$

Thus the average FEXT power from one source with bipolar signalling is

$$Q_F(1) = -M_F + 10 \log \int_0^{2f_o} |R(f)|^2 \cdot \frac{4f_o}{Z_o} \left(\frac{f}{f_o}\right)^2 x$$

$$\text{Sin}^2\ 2Bdf \qquad \text{------------------} \qquad (B.2)$$

It is seen that this power is independent of the repeater section loss and depends only on channel shaping.

If $D_F(S)$ dBW is the normalized FEXT interference (i.e. when $M_F = 0$, and $\frac{V_r^2}{Z_o} = 1$, $Q_F(1) = D_F(S)\}$, then

$$Q_F(1) = -M_F + 10 \log \frac{V_r^2}{Z_o} + D_F(s) \text{ --------------} (B.3)$$

$D_F(S)$ for raised-cosine shaping has been evaluated (6), by numerical integration to give:

$$D_F(RC) = 10 \log \int_0^{2f_o} \frac{1}{f_o} \left(\frac{f}{f_o}\right)^2 \sin^2 2B \cos^4 B \, df$$

$$= -7.4 \text{ dB}.$$

As for NEXT, with n systems, the overall FEXT interference is

$$Q_F(n) = M_F + D_F(S) + 10 \log \frac{V_r^2}{Z_o} + I(\sigma_N, n)$$

$$+ 2.33\sigma_\phi (\sigma_N, n) \text{ --------------- } \quad (B.4)$$

For 99% of the repeaters to have a worst error of 10^{-7}, then a S/N ratio larger than 20.8 dB is required

or $M_F - D_F(S) - I - 2.33\sigma_\phi \geq 20.8 + M_\varepsilon$.

For raised cosine shaping and correcting for effects of HDB3 coding,

$$M_F \geq 13.76 + I + 2.33\sigma_\phi + M_\varepsilon.$$

For any other shaping,

$$M_F \geq 13.76 + I + 2.33\sigma_\phi + \Delta_F(S) + M_\varepsilon \text{ -----------} \quad (B.5)$$

Where the correction factor $\Delta_F(S)$ = 1.2 dB for cosine shaping and $\Delta_F(S)$ = -0.3 dB for Gaussian shaping (6).

APPENDIX C

Central Office Noise

The operation of electromagnetic relays in telephone offices produces electrical transients which propagate down the pairs. Through crosstalk in the cable, these disturbances may appear on pairs not connected to switches. The energy distribution of this noise is very broad, reaching frequencies in excess of 2 megacycles. The transients are of several types and are quite complex. Since the noise is primarily due to switching, the error-rate is strongly dependent upon office activity, being highest during the busy hours and disappearing in the early hours of morning. Since a crosstalk path is involved, the levels of the noise transients reflect the approximately log-normal distribution of crosstalk loss and vary widely.

Because of the complexity of the office noise phenomenon, it is necessary to lean heavily on experimental results. The feature of the noise which is of most interest in PCM transmission is the error rate which it produces in the office repeater as a function of the length of the repeater section. As before, the repeater may be assumed to make an error if the S/N ratio at the decision point is less than

20.8 dB. The error rates to be expected on typical office repeater sections have been studied through extensive measurements (7) and the following general conclusions may be used:

(a) The signal-to-noise ratio, and hence error rate of the office repeater is determined by the difference between two losses. The first of these is the loss in the path by which the noise reaches the repeater input. It is convenient to imagine that the noise originates at the office termination of the switched voice-frequency pairs (see fig. 3.3). Thus, the noise path loss is the sum of a crosstalk coupling loss and a direct transmission loss. The typical mean near-end coupling loss is 76.8 dB at 1024 KHz, while in the case of far-end coupling the loss may be approximated by

$$M_F \simeq 65.5-10 \log (\ell/1862) \text{ ----------------- (C.1)}$$

where M_F = mean equal-level far-end crosstalk coupling loss at 1024 KHz for a cable of length ℓ metres. For the direct transmission loss, it is adequate to substitute the attenuation at 1024 KHz of the pairs over which the noise travels. The second loss needed, is that in the signal path and this is simply the cable loss incoming to the repeater. Under very severe noise conditions, an error rate of 10^{-7} is attained if the difference between the noise path loss and signal path loss is at least 52 dB.

(b) In repeater sections near offices, office noise and crosstalk are both present. It has been found that the error rate due to the sum of the two types of interferences, is approximately the sum of the error rates due to the two interferences separately. For practical purposes, design requirements for office noise and crosstalk may therefore be considered separately.

APPENDIX D

Power Spectral Density of the Line Code

It can be shown (4), that the power spectral density P(f) of a bipolar pulse train with uncorrelated pulses and spaces is given by:-

$$P(f) = \frac{2p(1-p)}{T} \, |G(f)|^2 \, \frac{1-\mathrm{Cos}\ 2\pi f\ T}{1+2(2p-1)\mathrm{Cos}2\pi fT+(2p-2)^2} \quad \text{-----------------} \quad (D.1)$$

where G(f) = Fourier transform of the pulse shape

p = probability of occurrence of a "one"

1-p = probability of a "zero"

T = pulse repetition period.

For balanced binary data i.e. $p = \frac{1}{2}$, (D.1) simplifies to

$$P(f) = \frac{1}{T} \, |G(f)|^2 \mathrm{Sin}\ \Pi fT \quad \text{-----------------} (D.2)$$

If the pulses have width equal to T i.e. 100% duty cycle pulses, the Fourier transform of the pulse shape is given by (4)

$$G(f) = VT \, \frac{\mathrm{Sin}\ \pi fT}{\pi fT} \quad \text{-------------------} (D.3)$$

Substituting for G(f) in (D.2), the power spectral density for a bipolar pulse train with 100% duty-cycle rectangular pulses is given by:

$$P(f) = V^2 T \left| \frac{\text{Sin} \ \Pi fT}{\Pi ft} \right|^2 \text{Sin}^2 \pi \ fT \text{ ------- (D.4)}$$

However, if half width rectangular pulses i.e. 50% duty-cycle pulses as those shown in fig. D.1 are used, there is a reduction in the power spectrum of the pulse stream. In this case the Fourier transform of the pulse shape is given by:

$$G(f) = V \frac{T}{2} \quad \frac{\text{Sin} \ \pi fT/2}{\pi fT/_2} \text{ ------- (D.5)}$$

So that the power spectral density of 50% duty-cycle pulses is

$$P(f) = \frac{V^2 T}{4} \left| \frac{\text{Sin}\pi fT/2}{\pi fT/_2} \right|^2 \text{Sin}^2 \pi fT \text{ ----(D.6)}$$

Comparing equations D.4 and D.6, it is clear that NEXT and FEXT interference, which are proportional to the power spectral density of the transmitted pulse stream, are reduced by a factor of four if 50% duty-cycle pulses are used instead of 100% duty-cycle pulses. For this reason and the fact that with 100% duty-cycle pulses complete retiming and pulse width control in the regenerative repeater are difficult, 50% duty-cycle pulses are used in the 32-channel PCM system and most of the other PCM systems.

Fig. D.2 shows the normalised power spectral density of a bipolar code with half-width pulses. It has maximum power at half the bit rate (1024 KHz for the 32-channel PCM) and nulls at zero frequency and multiplies of the signalling frequency.

<u>Effect of HDB3 code on the Power Spetral Density</u>

The HDB3 code is formed:

By replacing 4 or more consecutive zero in the bipolar stream by one of the following sequencies:

BOOV

or OOOV

where B is a positive or negative pulse generated using the bipolar rule

and V is a pulse inserted or inverted in the original sequence and violates the bipolar rule. The choice of either sequence must ensure that the number of B pulse between consecutive V pulses is odd so that V pulses alternate in polarity.

Example:

Bipolar stream 100000110000001

HDB3 stream B000V0BB B00V0

Polarity of Pulses +000+0-+-00-0-

The underlined patterns are those inserted in the bipolar code using the HDB3 rule.

Extraction of long strings of zeros from the bipolar stream improves the timing content of the code but at the expense of increased power. It has been shown (16, 18) that the HDB3 code increases the power by a factor of 1.087 over the bipolar code at half the bit rate. Fig. D.3 compares the power spectra of the bipolar and HDB3 codes.

APPENDIX E

Cable Structure

The KP & T cables under investigation are of the paper-core star-quad type and their main physical features will be covered briefly in this section.

Fig. E.1 shows how the individual conductors are first arranged in groups of four. Each conductor may have a diameter as large as 1.3 mm, but 0.9 mm and 0.65 mm diameter conductors may be used in lighter cables. Over each conductor is wound a helix of paper string. This acts as a spacer and holds the paper tape lapping which is wound on top, away from the surface of the conductor. As a result a large percentage of the volume of the insulation between conductors is air - a most efficient of dielectrics. A cotton whipping holds four conductor units together round a central paper string to form a quad. In one quad, the two conductors, diagonally opposite each other, form a pair, i.e. they are used as the two conductors of one circuit.

Fig. E.2 shows how a number of quads is arranged in layers to form a complete cable. For simplicity, only two layers are shown but three, four or more may be used depending on how many circuits are required. The centre core or innermost layer may have 1, 3 or 4 quads. The number in each succeeding layer is then 6 greater than the number in the previous layer. In fig. E.2, a layer of 4 and a second layer of 10 are shown. A third-layer would consist of 16 quads. Layers are separated by a winding of cotton. Surrounding the outside layer are two paper layers and an outer sheath of lead.

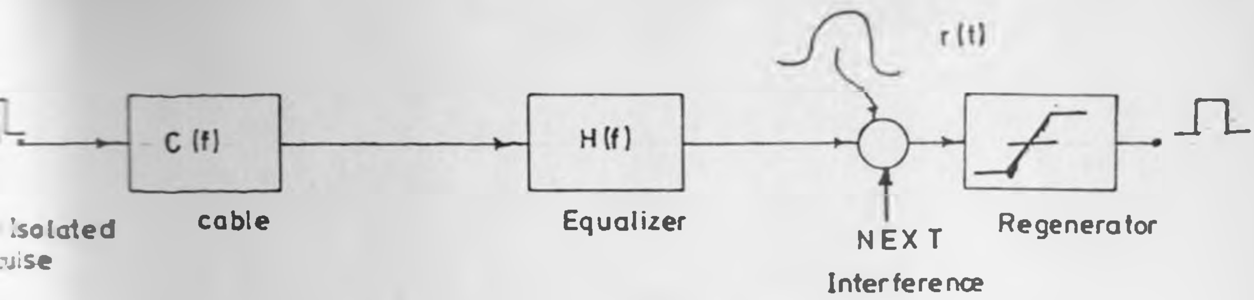


Fig A1 System model for NEXT

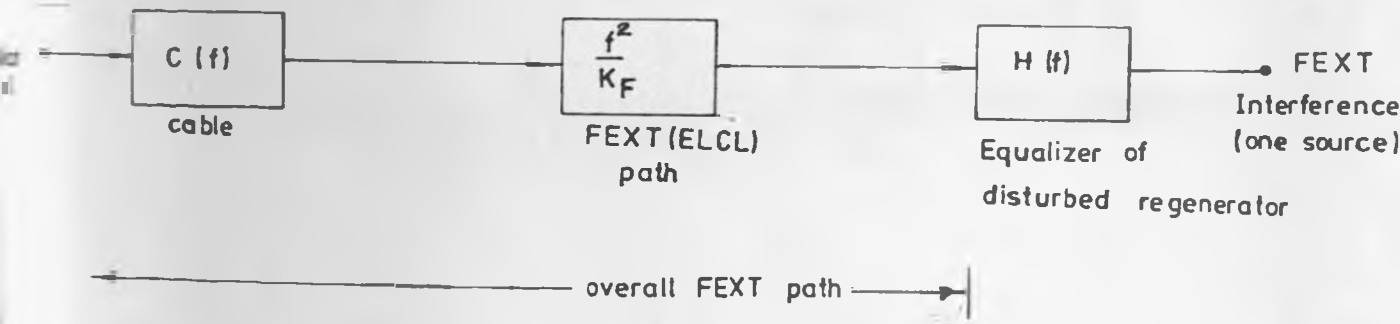


Fig B1 System model for FEXT

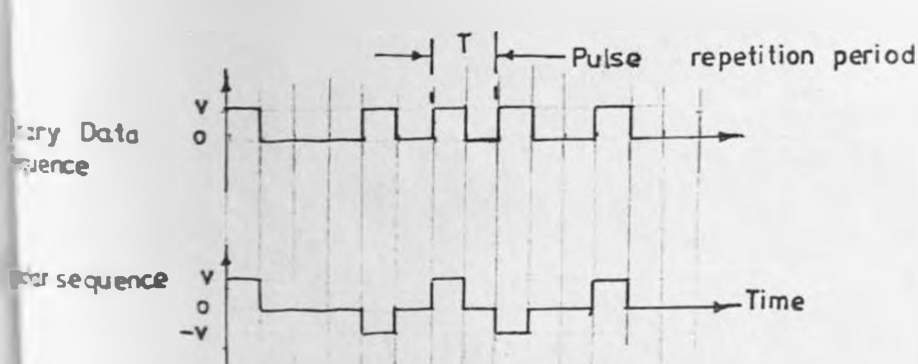


Fig D1 Bipolar sequence (with half width rectangular pulses) derived from binary data sequence

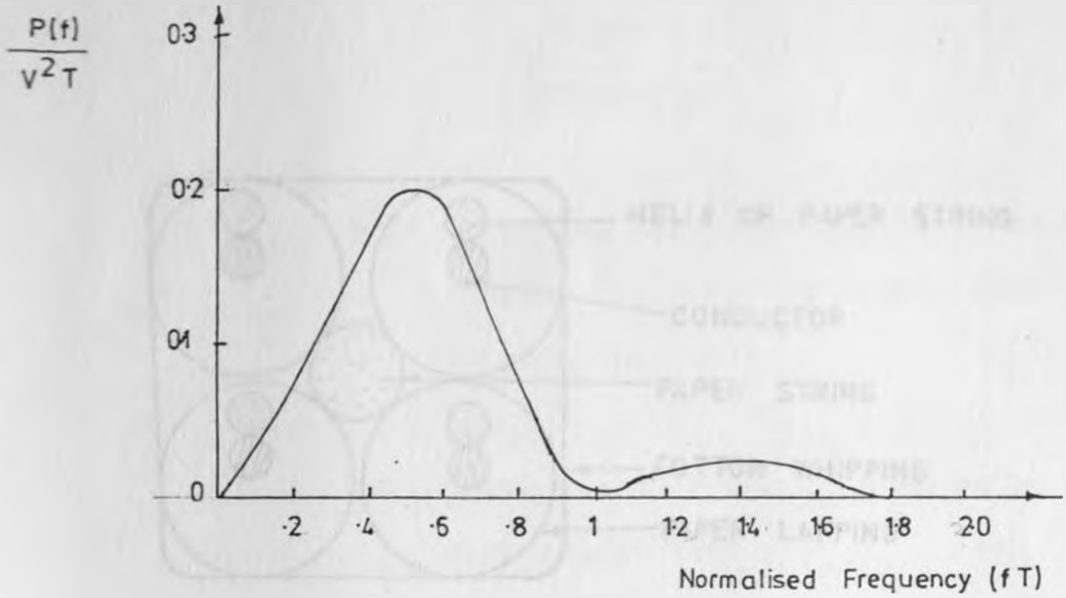


Fig D 2 Power spectral density of a random bipolar sequence with half — width rectangular pulses

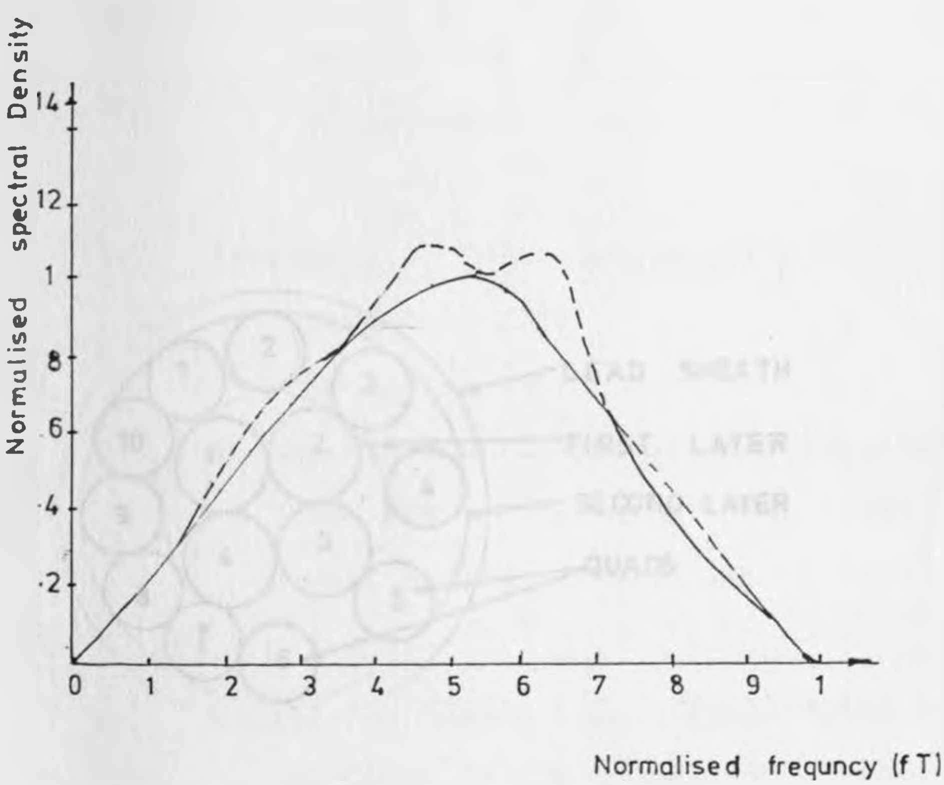


Fig D3 Power spectrum comparison of bipolar and HDB 3 Codes

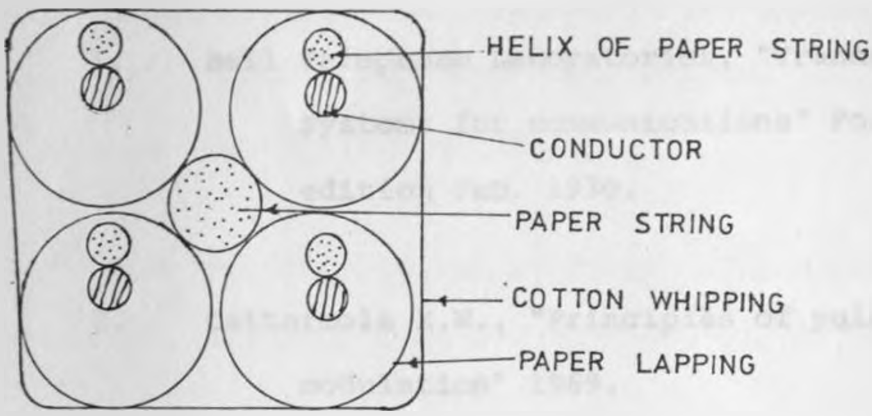


Fig E1 The cross section of a paper-insulated quad

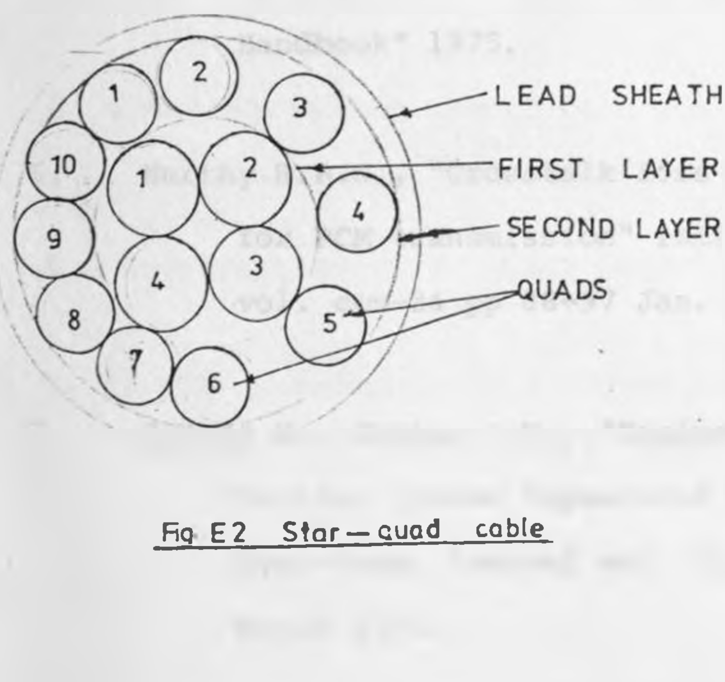


Fig E2 Star-quad cable

REFERENCES

1. Bell Telephone Laboratories, "Transmission Systems for communications" Fourth edition Feb. 1970.

2. Cattermole K.W., "Principles of pulse code modulation" 1969.

3. G.C. Hartley, et. al., "Techniques of Pulse-code modulation in communication networks" 1967.

4. Bennett W.R., Davey J.R., "Data Transmission" 1965.

5. Freeman R.L. "Telecommunication Transmission Handbook" 1975.

6. Murthy B.R.N., "Crosstalk Loss Requirement for PCM transmission" IEEE Trans. Commun. vol. com-24 pp 88-97 Jan. 1976.

7. Cravis H., Grater T.V., "Engineering of Tl Carrier System Repeatered Lines" Bell Syst-Tech. Journal vol. 42, pp 431-486 March 1973.

8. Aaron M, "PCM Transmission in the Exchange Plant" Bell Syst Tech Journal vol. 41 pp 99-141 Jan. 1962.

9. BTM-ITT, "PCM System Application Handbook" 6th Edition, Part 3 System Crosstalk Performance.

10. Selected Articles from the "Lenkurt Demodulator" v.3 CTE Lenkurt:-

    (1) PCM Repeatered Lines P. 85

    (2) PCM Field Study P. 101

    (3) PCM Cable Considerations P. 109

    (4) Cables Tests and Measurements for PCM.

11. Plessey, Transmission, "Cable and Pair Testing"

12. Telettra PCM System Description and Installation.

13. PCM Line Test Set ZTTR 00102 Description.

14. C.C.I.T.T. Recommendations 700 Series - Geneva 1972, Ammended at Geneva 1976.

15. "An Experimental PCM System for Short Haul Trunks" B.S.T.J. 41 Jan. 1962 Pl. Davis C.G.

16. Duc N.Q., "Line Coding Techniques for Base-band Digital Transmission" Australian Telecommunication Research (ATR) Vol. 9 No. 1 1975.

17. Duc N.Q., Smith B.M., "Line Coding for Digital Data Transmission" ATR Vol. 11 No. 2 1977.

18. Croisier A., "Compatible High-Density Bipolar Codes: An unrestricted Transmission plan for PCM carriers" IEE Trans. Comm. Tech. (concise Papers) June 1970 Vol. Com-18 pp 265-268.