

# BBC

## ENGINEERING DIVISION

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### Pulse-code modulation for high-quality sound-signal distribution

by

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(Research Department, BBC Engineering Division)

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BRITISH BROADCASTING CORPORATION

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PULSE-CODE MODULATION FOR HIGH-QUALITY  
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## FOREWORD

**T**HIS is one of a series of Engineering Monographs published by the British Broadcasting Corporation. About six are produced every year, each dealing with a technical subject within the field of television and sound broadcasting. Each Monograph describes work that has been done by the Engineering Division of the BBC and includes, where appropriate, a survey of earlier work on the same subject. From time to time the series may include selected reprints of articles by BBC authors that have appeared in technical journals. Papers dealing with general engineering developments in broadcasting may also be included occasionally.

This series should be of interest and value to engineers engaged in the fields of broadcasting and of telecommunications generally.

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# PULSE-CODE MODULATION FOR HIGH-QUALITY SOUND-SIGNAL DISTRIBUTION

## SUMMARY

The possibility is considered of employing pulse-code modulation (p.c.m.) signals, carried over wide-band links, to distribute the sound signals from radio or television programme sources to the various transmitters and the use of a Post Office p.c.m. network for this purpose is discussed. Experimental equipment has been constructed for coding and decoding broadcast-quality sound signals.

One of the applications of p.c.m. envisaged for the near future is the incorporation of the sound signal associated with a television programme into the video waveform for distribution over existing links. Suitable equipment for combining and separating sound and vision signals at the sending and receiving ends of a link has been constructed.

It is concluded that p.c.m. systems can now be designed to meet all the requirements, both technical and operational, of a high-quality sound distribution network.

## 1. Introduction

Pulse-code modulation (p.c.m.), which may be regarded as a kind of high-speed telegraphy, is a very attractive method of distributing sound signals from a broadcast studio or other programme source to the various transmitters. The audio-frequency output can be converted into corresponding p.c.m. signals, which consist of pulses arranged in groups according to a prescribed code; these pulses can be detected and decoded even in the presence of considerable noise and distortion on the distribution network. A number of different programmes can be distributed over a common link by interleaving the corresponding pulses in time (time division multiplex or t.d.m.) without introducing crosstalk, while individual programme signals can easily be introduced or withdrawn as required at any point in the network. Finally, the t.d.m. principle could be applied to the distribution of the sound signal associated with a television programme by introducing p.c.m. signals during the line blanking interval.

Although the application of pulse-code modulation to telephony has already been dealt with very fully in the literature<sup>1,2,3,4</sup>, the special problems which arise in the use of this technique for the point-to-point distribution of broadcast programme material have not received much attention. It was, therefore, decided to undertake a programme of work to determine whether p.c.m. can be used for this purpose in the immediate future and to specify the optimum parameters of a suitable system; the purpose of the present monograph is to give an account of this investigation.

## 2. Basic Concept of Pulse-code Modulation

### 2.1 Sampling and Coding

Pulse-code modulation, like all time-division-multiplex systems, is based on the sampling theorems expounded by Nyquist and Shannon<sup>5,6</sup>. These theorems may be summarized in a simplified form as follows: in order to transmit any waveform as a sequence of discrete amplitude-samples, it is necessary to send at least two samples per cycle of the highest frequency Fourier component of the

waveform. For example, in order to transmit a sound signal which has a maximum frequency component of 15 kHz, it is necessary to sample at a rate somewhat greater than 30 kHz.

The signal which results from an amplitude sampling process is said to be pulse-amplitude modulated (p.a.m.) and it can be transmitted to the receiving terminal in any form thought desirable or convenient. At the receiving terminal the original waveform can be reconstructed from the p.a.m. signal by generating a  $\sin x/x$ -shaped pulse proportional in amplitude to the sample pulse and having its zeros at those instants at which adjacent pulses have their maxima; the output signal is obtained by arithmetically summing these  $\sin x/x$  pulses.

Fig. 1 shows in block diagram form a typical sampling arrangement as used in any t.d.m. system. The input signal is band-limited by a low-pass filter to ensure that it is not possible to exceed the maximum frequency permitted by the sampling rate; the resulting signal is then sampled to produce a p.a.m. signal which is coded into the form chosen for transmission (pulse-position modulation, pulse-duration modulation, pulse-code modulation, etc.). At the receiving terminal the received signal is decoded into a p.a.m. form and the original waveform is reconstructed by means of a second low-pass filter, which in effect adds appropriate  $\sin x/x$  pulses.\* In practice it is necessary, in order to secure adequate attenuation of unwanted components, to make the cut-off frequency of the low-pass filters somewhat less than half the sampling frequency.

In p.c.m., the magnitude of the sample is coded into a digital number which in turn is transmitted as a group of pulses. Each pulse may be made to assume one of a number of predetermined states—for example, the amplitude may have one of several discrete values or the polarity may be reversed—and means are provided at the receiving end of the system for distinguishing, without ambiguity,

\* To obtain distortionless transmission of the waveform, both the input and the output low-pass filters must approximate to the ideal. For sound transmission, however, because of the relative insensitivity of the ear to moderate phase distortion<sup>7</sup>, it is permissible in practice to use conventional sharp-cut filters without group-delay correction.

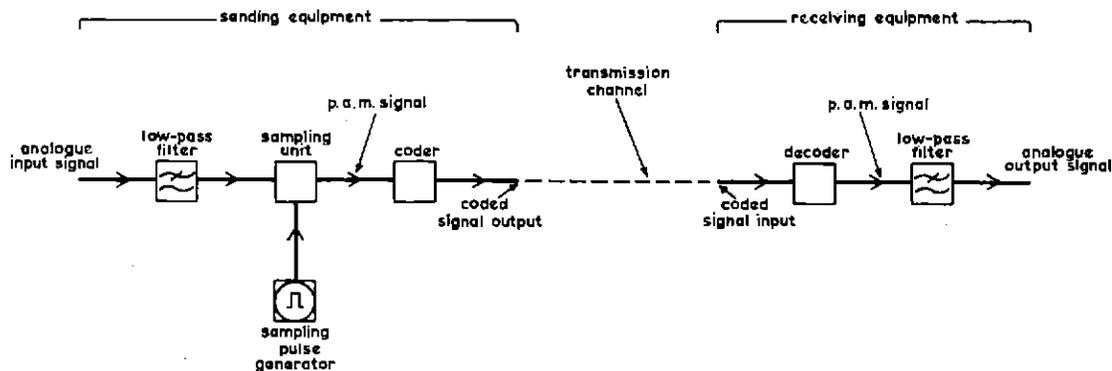


Fig. 1 — Sampling arrangement for a t.d.m. system

between one state and another. The number of alternative states which each pulse may assume is referred to as the base of the code.

One of the advantages of p.c.m. is that the form of coding used may be adapted to suit the characteristics of the communication channel over which the pulses are to be transmitted. The code is characterized by the number of pulses,  $n$ , corresponding to each signal sample and the number of states,  $m$ , which each pulse may assume. The communication channel, on the other hand, can be described, in essence, by its bandwidth and its signal-to-noise ratio. From the channel bandwidth can be deduced\* the minimum spacing at which pulses can be transmitted and received without mutual interference, and thence, for a given sampling frequency, the maximum value of  $n$ . The signal-to-noise ratio sets an upper limit to the value of  $m$ , because the effect of noise is to introduce ambiguity in distinguishing between different states—for example, between different pulse amplitudes. The rate at which information is transmitted is proportional to  $m^n$ , and it is thus possible, by an appropriate combination of  $m$  and  $n$ , to make the most efficient use of the available channel bandwidth and signal-to-noise ratio by trading, in effect, the one quantity for the other. It should further be noted that pulse-coded signals can, if necessary, be transformed from one base to another at any point in the communication system by means of appropriate logic circuits, the rate of transmission of information remaining unchanged.

The most salient feature of p.c.m. is its extreme robustness in transmission. Noise added to the coded signals in course of transmission, unless of sufficient magnitude to cause errors in decoding, has no effect on the output signal. If the degree of pulse distortion or noise introduced in the full length of the communication channel is such as to prevent unambiguous detection of the coded signals, the latter can be detected and completely regenerated at one or more intermediate points; thus, the impairments produced by a p.c.m. system can be confined to those arising in the coding and decoding operation and do not increase with the distance over which the information is sent.

The commonest form of p.c.m. employs a binary code,

\* With due allowance for the amplitude/frequency and phase/frequency characteristics of the system within the working band.

i.e.  $m$  is equal to 2, the information being conveyed by the presence or absence of a pulse at a given instant of time. This type of coding is suitable for communication channels having a signal-to-noise ratio several orders of magnitude less than would be required for direct (i.e. analogue) transmission of the incoming signal; in fact it can be shown<sup>9</sup> from statistical considerations that with random noise having an r.m.s. value only 23 dB below the pulse amplitude, the rate of error amounts to one pulse in  $10^{12}$ , which can probably be regarded as negligible (the channel bandwidth must, of course, be sufficient to compensate for this deficiency in terms of information-carrying capacity). Since the decoding process is not critically dependent on the amplitude of the received pulses, non-linear distortion in the channel has little or no effect.

The binary-coded p.c.m. system, which was invented in 1937<sup>8</sup>, has been exhaustively treated in the literature and only a brief résumé of its salient features will be given here. Although the remainder of this section is written in terms of a binary-coded signal, much of this material applies equally to transmission by codes having bases other than 2.

Fig. 2 illustrates the process of coding and decoding in a 4-digit binary p.c.m. system, in which the instantaneous signal level is represented by any one of sixteen different codes. The significance of this latter process is discussed in the next section.

## 2.2 Quantizing Noise

The principal disadvantage of p.c.m. is caused by the fact that, as the amplitude of each sample is described by a digital number, the data to be transmitted must be represented by a set of discrete amplitudes; the output obtained from a p.c.m. system is therefore quantized into discrete levels, the number of which is determined by the number of digits per sample. The output signal can thus be regarded as a perfect representation of the input signal together with an error caused by the quantizing. It is self-evident that the magnitude of this error lies within the range of plus and minus half of one quantizing step, but its subjective effect is not so obvious. The audible effect of the error produced by quantizing closely resembles that of random (white) noise provided that the number of quantizing levels covered by the signal is not too small<sup>4</sup>; for this reason the error is always referred to as 'quantizing noise'.

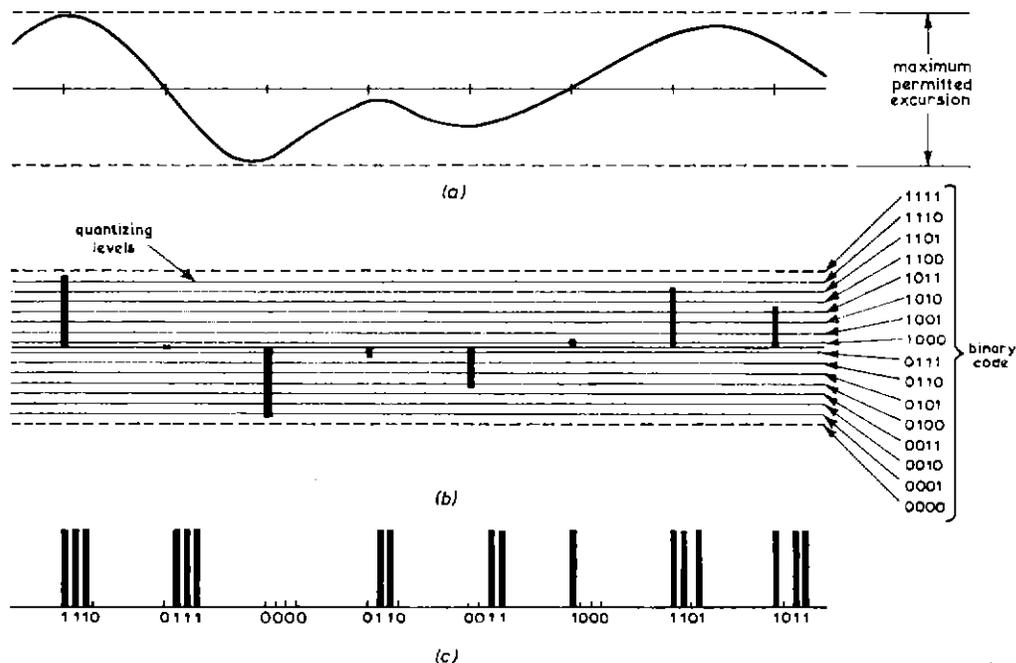


Fig. 2 — Binary p.c.m. coding for a 4-bit system

- (a) Band-limited analogue signal.  
 (b) Pulse amplitude modulated signal derived from (a); the binary code used to represent the quantizing levels is shown on the right.  
 (c) Binary pulse-code modulated signal derived from (b).

The calculated values of signal-to-quantizing noise ratio for 10-, 11-, and 12-digit p.c.m. systems are shown in Table 1; it is assumed that the quantizing steps are equally spaced. To attain these figures in practice, a high order of instrumental accuracy is required. This point is illustrated in the last column of Table 1, which shows the percentage of the signal amplitude contributed by the least significant digit in each case; these figures give an indication of the accuracy with which the amplitude of a sample must be measured at the sending terminal and reconstructed at the receiving terminal if the least significant digit is to be effective.

Subjective tests using white noise, the amplitude of which had a probability density of gaussian form, have shown that in a distribution system for high-quality sound signals, the ratio of peak signal to r.m.s. noise should be at least 72 dB. Quantizing noise can be shown to differ from gaussian noise in that the amplitude has a rect-

angular probability density. If, however, it is assumed that quantizing noise and Gaussian noise of equal r.m.s. value sound equally loud, it appears from the table that a p.c.m. system using 10 binary digits (bits) per sample would be unsatisfactory, 11 bits per sample would be on the threshold of acceptability, while a 12-bit-per-sample system would probably be adequate.

In the complete absence of signal, quantizing noise naturally ceases. In practice, a small amount of incoming noise, or interference such as hum, is likely to be present at the input to the p.c.m. system, and if this is sufficient to cause the analogue-to-digital converter to jump intermittently from one quantizing level to the next, corresponding voltage steps will appear at the receiving end of the system, producing idling noise. Because of its discontinuous nature, idling noise is more objectionable to the ear than the background of continuous hiss produced by quantizing noise; the distinction, however, becomes

TABLE 1  
 Signal-to-quantizing Noise Ratio in Binary P.C.M. Systems

Number of digits	Number of quantizing levels	$\frac{\text{Peak signal}}{\text{r.m.s. quantizing noise}}$	Contribution of least significant digit
10	1,024	65 dB	per cent 0.1
11	2,048	71 dB	0.05
12	4,096	77 dB	0.025

less at low sound levels, and it has been found that if quantizing noise can be so far reduced as to be barely audible, idling noise can usually be neglected.

The assumption that quantizing noise is subjectively equivalent to white noise is valid only when the signal extends over a large number of quantizing levels. Whenever the signal amplitude falls so low that only the first few of the available levels are utilized, the subjective effect of quantizing begins to resemble non-linear distortion rather than random noise, while with still lower amplitudes, interruptions in the sound may be heard when the signal momentarily fails to reach the first quantizing level. In a high-quality sound system, in which the signal covers a wide dynamic range, care must be taken that such effects are inaudible even in the quietest passages

The quantizing noise at the output of a p.c.m. system will be increased if the signal is repeatedly decoded and recoded; the increase in noise level will be 3 dB for one additional decoding and recoding operation, 4.8 dB for two such operations, 6 dB for three, and so on. However, there is no fundamental reason why a signal need be coded and decoded more than once; the extraction of one or more out of a number of programmes carried by a 'bit stream' is a simple operation as is also its insertion. The mixing of signals representing programme contributions from several sources can be effected by the use of digital techniques similar to those employed in computers.

### 2.3 Compatibility with Existing P.C.M. Systems

The transmission channels likely to be used for programme distribution by p.c.m. have an information capacity greatly in excess of programme requirements and for efficient utilization must be shared by a number of services. It is therefore desirable that the standards adopted for programme distribution should be compatible with those adopted by the Post Office; even if the BBC were to operate a separate distribution network, it would be advantageous to be able to employ commercial equipment of the type developed for Post Office use.

The circuits envisaged by the Post Office<sup>9</sup> are based on a sampling rate of 8 kHz, 8 bits per sample ( $m=2$ ,  $n=8$ ) and 24 channels per circuit; the bit rate for a standard Post Office circuit will therefore be 1,536k bits/sec ( $8k \times 8 \times 24$ ). It has already been stated that a sampling rate in excess of 30 kHz is necessary for signals of broadcast quality, which suggests that a convenient choice of parameters for the efficient use of a standard circuit might be based, for example, on four monophonic or two stereophonic programmes, utilizing a sampling rate of 32 kHz, each sample consisting of 12 bits. In this case, however, the coded samples would occupy the entire channel capacity leaving no room for synchronizing signals; it would then be necessary to use sophisticated statistical methods to decode the received signals correctly<sup>10</sup>. In a practical coding system it is usual to send the digit pulses representing one sample from each of the channels sequentially, and a block of pulses which represents a total of one sample from every channel is known as a frame. In order to decode the signals it is necessary to determine which of the pulses in the se-

quence is the first in the frame and adjust the logic circuits of the decoder accordingly, a process known as framing.\* For this reason it might be necessary, in the case under consideration, to restrict the number of bits per sample to 11, thus leaving 4 bits per frame, i.e. 128k bits/sec ( $4 \times 32k$  bits/sec), for other purposes. For framing purposes 32k bits/sec would be sufficient, leaving 96k bits/sec which could be used for a telephony channel (64k bits/sec) plus monitoring or teleprinter facilities (32k bits/sec.)

The above calculation does not take into account the possibility that, in order to conform with existing p.c.m. transmission systems, it may, in practice, be necessary to place some restriction on the codes that may be used—for example, by excluding certain combinations of digits or by introducing redundancy<sup>11</sup>. In these circumstances, it may be that the whole of the information capacity of a 1,536k bits/sec channel will be required for the transmission of four 11-digit programme signals.

### 2.4 Practical Experience

In view of the foregoing considerations, it was decided as a feasibility study, to construct in the laboratory a p.c.m. system to the following specification:

Sampling frequency:	32 kHz
Bits per sample:	11
Upper a.f. limit:	14 kHz (half sampling frequency less allowance for imperfect filters)
Lower a.f. limit:	determined by associated a.f. terminal equipment

Experience gained with this equipment showed that it is possible to obtain a signal-to-noise ratio equal to the theoretical figure given in Table 1 within the limits of experimental error ( $\pm 1$  dB). Techniques which will be referred to later can be used to improve the signal-to-noise ratio of a simple p.c.m. system; it therefore seems likely that an acceptable system based on 11 or even 10 digits per sample could be achieved.

## 3. Coding and Decoding

### 3.1 General

It has been pointed out that the process of quantizing a continuum of input signal values into a finite set of codes implies coding errors which manifest themselves as noise.

In theory, adding an extra binary digit to the code, thereby doubling the number of quantizing levels and

\* At the sending end of the system, successive frame pulses are switched on or off according to a predetermined time pattern—for example, it may be arranged that every other pulse is absent. At the receiving end, an identical time pattern is generated, and compared, over a suitable time interval, with the pattern of received pulses appearing in a series of 'time slots' one frame-interval apart. If the system is out of synchronization, the time slots which should contain the framing-pulse pattern will be occupied instead by some fortuitous sequence derived from the signal pulses; the probability that this latter sequence will be the same as the framing-pulse pattern and remain so while the comparison is being made is extremely small, and can be reduced to negligible proportions by increasing the number of time slots to be compared. Correction for the out-of-synchronization condition is made by causing the decoder timing to slip by one digit at a time until the prescribed framing-pulse pattern emerges.

halving the separation between successive levels, should reduce the quantizing noise by 6 dB. In practice the expected reduction in noise level with each additional bit in the code is achieved only up to the point at which inaccuracies in the converters become appreciable. The most obvious cause of inaccurate coding and decoding arises from errors in the analogue circuits.

A less obvious but equally important cause of inaccuracy in decoding equipment is timing error. Jitter on the timing of the samples taken by the coder causes an error proportional to the rate of change of the input signal. Timing jitter in the decoder which modulates the width and/or position of the amplitude-modulated pulses at the output of the decoder will also produce noise components in the reconstituted signal. For example, in a p.c.m. system sampling at 30 kHz, coding an 11-bit word, a pulse width error of 16.3 ns in the output of the digital-to-analogue converter would produce a noise voltage equal to the contribution to the signal of the least significant bit.

Because of the possibility of errors within the coding and decoding equipment it is necessary to measure the level of noise in the recovered analogue signal and by comparison of the theoretical and measured values establish the performance of the equipment.

Quantizing noise exists only in the presence of a signal, and therefore special techniques are needed in order to measure it. One method relies upon the use of a signal whose spectrum occupies a restricted portion of the base bandwidth. The quantizing noise produced within that portion of the base bandwidth which is not occupied by the signal is then measured. In another method thermal noise injected into a parallel reference circuit having the same gain as the p.c.m. chain is adjusted by ear, in the presence of a programme signal, to be equal in loudness to the quantizing noise. The level of the injected noise in the reference chain is then measured in the absence of signal. It has been found that the adjustment of injected noise to equality with the p.c.m. noise can be easily repeated to within  $\pm 1$  dB.

The remainder of this section is concerned with the basic principles of coding and decoding and with the development of equipment suitable for use with high-quality sound signals.

### 3.2 Analogue to Digital Conversion

#### 3.2.1 Methods of conversion

A large number of analogue-to-digital conversion methods is known but all derive from three basic methods according to the way in which the amplitude of the signal sample is measured.

Converters of the first kind compare the voltage of the signal sample with a number of precise voltage levels presented successively in descending order of magnitude; each level corresponds to the magnitude signified by one digit position in the code. As soon as a level is reached which is less than the voltage of the sample, subtraction takes place and a '1' is generated for that digit position; the difference signal is then dealt with in the same way and the process is repeated until all the digit positions have been explored.

This method is known as successive-approximation conversion.

In converters of the second kind known as the 'counter' type, the amplitude of the sample is measured by counting out the number of units contained in it until less than one remains. This may be done by comparing the signal sample with a uniformly rising ramp; during the ramp period a counter runs at constant speed and accumulates a number proportional to time. When the ramp voltage equals the sample voltage the counter is stopped and the converter output read from the counter.

In the third kind of converter, known as a 'parallel converter', the sample amplitude is measured by comparing its voltage simultaneously with the complete set of possible quantized values. The coding of high-quality sound signals requires high incremental accuracy; this requirement eliminates parallel-converters because of the complexity and expense of a circuit containing the extremely large number of comparators which would be required.

In certain respects the requirements for audio-signal transmission are far less stringent than those for, for example, data transmission. For audio signals, slowly varying errors of signal amplitude up to about  $\pm 5$  per cent are tolerable, and the fact that the audio-frequency band does not extend to zero frequency implies that direct current injected into the signal chain as a by-product of coding and decoding is unimportant.

To produce the most economic design of a p.c.m. system for audio signals the relaxations above should be fully exploited and the coding and decoding methods adopted should of their very nature provide the theoretical quantizing noise level with the minimum of pre-set adjustments. In the counter method described above, the output of the converter is monotonically related to the input. No discontinuities can occur in the conversion law because no sets of current or voltage references requiring accurate adjustment are used, an advantage shared by no other coding method.

For these reasons the counter method of coding was adopted for the sound-signal analogue-to-digital converter (a.d.c.) described below.

#### 3.2.2 The Design of a Counter-type Analogue-to-digital Converter

Fig. 3 shows a schematic of a counter-type a.d.c.; the clock rates shown are appropriate to 11 bits and a sampling rate of approximately 30 kHz. The input signal is sampled and held and applied to one input of a voltage comparator. The other input to the comparator is provided by a voltage across a capacitor which is being charged from a constant current source. In the initial state, up to the moment the sample is applied, the capacitor is short-circuited and the counter is held in the reset condition. The short-circuit across the capacitor and the reset signal on the counter are then simultaneously removed; the capacitor then begins to charge and the counter begins to count the pulses supplied by the 80 MHz clock. When the voltage across the capacitor equals the voltage of the applied sample, a pulse

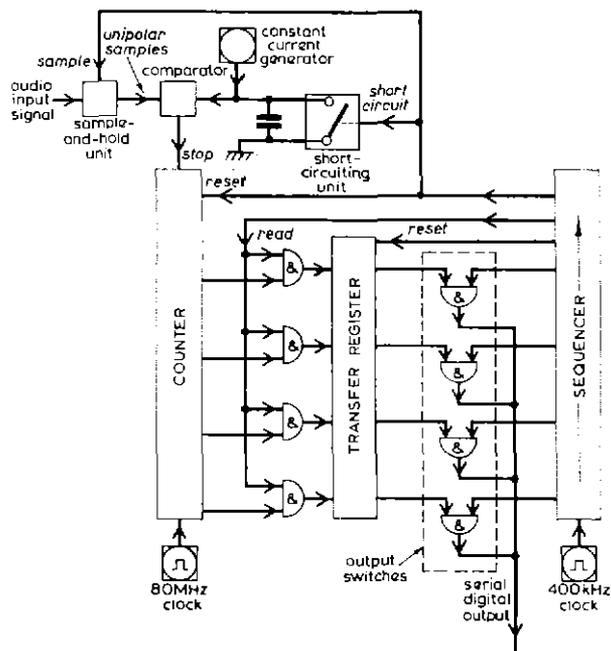


Fig. 3 — Counter-type analogue-to-digital converter block diagram (4 bits only shown for simplicity)

from the comparator stops the counter and the required digital representation of the input voltage sample is then available at the counter outputs.

The time interval between successive samples at 30 kHz sampling rate is  $33.3 \mu\text{s}$ . The counting process described above occupies a period of time proportional to the number produced but in all cases must be complete sufficiently quickly to allow the digital output to be read and the counter and capacitor voltage to be reset in time for the next conversion. The need to provide an 11-bit digital output implies a maximum count of 2,048, which, with a counting rate of 80 MHz, will be complete in  $25.6 \mu\text{s}$ .

The remainder of the a.d.c. is concerned only with storing the digits, reading them out in the proper sequence for transmission, and resetting the circuits ready for the next conversion. The sequence of control operations in the coder is dictated by a sequencer driven by a 400 kHz clock, and this clock determines the 'bit-rate' or rate at which the pulses comprising the digital signal leave the a.d.c. The 400 kHz bit-rate clock requires to be synchronized with the 80 MHz counter-driving clock, since the 400 kHz clock controls the moment when conversion begins. The start of counting must always occur synchronously with respect to the 80 MHz clock, or a random error would occur, the peak-to-peak amplitude of which would equal one quantum level. In practice, therefore, the 400 kHz pulses are derived from the 80 MHz pulses by frequency division.

Propagation delays in the comparator result only in the addition of a constant number to the generated code and hence in the addition to the reconstituted signal of a d.c. component which is easily removed at the analogue output. Propagation delays in the short-circuiting device and

the release circuitry of the counter have a similar effect, as does imperfect short-circuiting of the capacitor, the result of which would be to leave a small initial charge.

Slow variation with time of the nominally constant current charging the capacitor will change the magnitude of the signal slightly. Both these errors are acceptable in moderate amounts. The critical requirement of monotonicity in the conversion is intrinsic in the coding method.

One instrumental requirement which became apparent during the experimental work was the necessity of preventing overrun in the counter of the coder at maximum signal excursion. If allowed to overrun the counter would suddenly revert to zero and then start to count up again. The result is a particularly violent overload condition in which peaks of signal are transferred to the opposite extreme of the signal range with aurally disastrous consequences. Logic circuits may be connected to the counter to prevent any digits of more than a particular significance from being changed after they have all become '1'. This artifice ensures that if the counter subsequently overruns, only digits of low significance can be removed.

### 3.3 Digital-to-analogue Conversion

#### 3.3.1 Methods of Conversion

Three essentially different methods of decoding are known. The first of these was described by Oliver, Pierce, and Shannon in 1948<sup>2</sup> and is applicable to binary code only. The binary number to be decoded is presented in reverse order, i.e. with the least significant bit first and the most significant bit last. The reception of each bit in turn by the converter causes a fixed charge to be injected into the capacitor of a parallel capacitor-resistor circuit, the time constant of which is so adjusted that the voltage across the capacitor decays to half its value in the time interval between successive bits. The capacitor voltage immediately after reception of the last (most significant) bit is then directly proportional to the binary number.

In the second method a time interval is counted out proportional to the digital number to be decoded. During this time interval a linear process, for example the charging of a capacitor by a constant current, is allowed to proceed. At the end of the time interval the accumulated quantity, e.g. the voltage on the capacitor, is proportional to the digital number being decoded. This is known as the counter method of digital-to-analogue conversion.

For the third method of decoding the digital number to be converted is first assembled in a holding register. When the number is complete a 'command' pulse applied to gating circuits causes each digit representing '1' to turn on a current proportional to the value signified by that digit. The sum of the currents is then proportional to the digital number. This method of decoding corresponds roughly to the parallel or 'simultaneous coding' process.

#### 3.3.2 Design of a Counter-type Digital-to-analogue Converter

The considerations leading to the adoption of a counter-type coder lead with equal force to the adoption of a counter-type decoder. The two instruments, although they

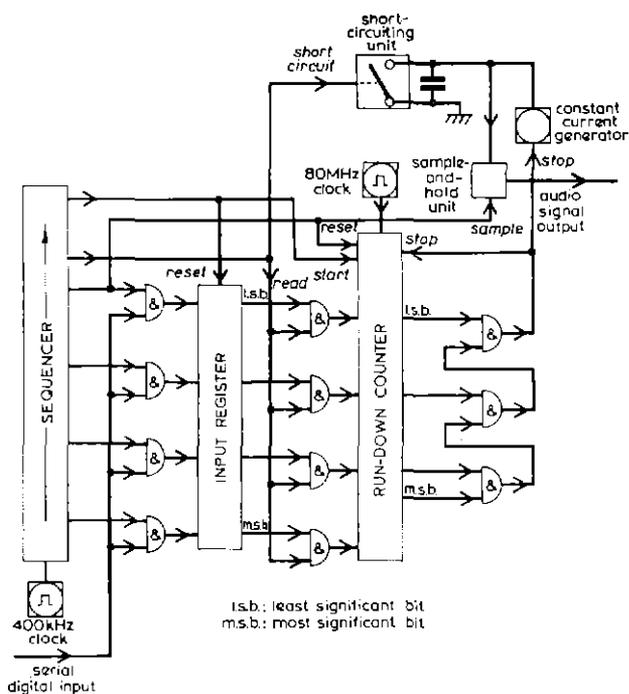


Fig. 4 — Counter-type digital-to-analogue converter block diagram (4 bits only shown for simplicity)

have much in common, differ in a number of respects. The design of a counter-type converter is described below.

Fig. 4 shows a schematic of a counter-type digital-to-analogue converter (d.a.c.) corresponding to the coder discussed in Section 3.2. Here the number to be decoded is set into a high-speed counter using presetting connections on the counting stages. After setting the counter, counting is started and a capacitor simultaneously begins to charge from a constant current source. The counter then counts down to zero under the control of an 80 MHz clock; the final transition to an all-zeros state, involving only the bistable stage for the least significant bit, stops the charging of the capacitor. The voltage on the latter is then proportional to the number being decoded and when sampled and filtered provides the analogue output.

The action of the counter d.a.c., as described in the simple explanation given above, is correct in principle but requires in practice a small modification to keep within the operating speed of currently available integrated circuits. The difficulty arises because the stop signal, as generated by the last bistable element in the counter, exists for only a very short time. Fortunately, by deriving the stop signal from the second or third bistable, its duration can be doubled or redoubled; thus, in a counter running at 80 MHz the duration of the stop signal can be increased from  $12\frac{1}{2}$  to 25 or 50 ns, a length of time sufficient to operate the counter-stopping circuits reliably. It should be noted that a penalty is incurred by the use of this artifice in that the counter will be stopped before it reaches zero. If the stop signal is derived from the 9th digit and 11 are provided, the counter will be stopped at the time when the number  $2^{(11-9)}-1$  remains. Thus in the case of an 11-bit decoder

the number of usable levels will be reduced from 2,048 to 2,045, which is negligible in practice.

The remarks made about tolerance of the audio signal to timing and capacitor starting voltage errors when discussing the counter-type a.d.c. are also applicable to the counter-type d.a.c.

### 3.4 Performance of Counter A.D.C. and D.A.C.

The ultimate limitation on the performance of a.d.c. and d.a.c. must lie with the analogue circuits involved; in counter instruments, in particular, it depends on the consistency and precision of the sample-and-hold, ramp, and comparator circuits as well as on the limit of discrimination of the comparator.

The signal-to-noise ratio of the present a.d.c. or d.a.c. connected in tandem, measured by the noise substitution method of Section 3.1, was found to equal the theoretical value within the limits of experimental error ( $\pm 1$  dB), indicating that the performance of the analogue circuits involved was adequate for 11-bit coding.

Measurement was made of the timing stability of the stop pulse from the comparator of the a.d.c. when coding repeatedly a steady applied voltage. The timing jitter visible on the stop pulse was approximately 3 ns. Since the counting rate for 11 bits is one count in 12.5 ns this indicates that the analogue circuits at present in use would be suitable for 12-bit coding, i.e. coding with a counting rate of 6.25 ns per count.

It will be seen from the above that the construction of counter-type instruments both for coding and decoding audio signals of broadcast quality presents no great difficulty. The counting frequencies involved are high by present-day standards, and are near to the fastest that can be achieved by modern integrated circuits. However, at the present rate of progress of integrated circuit performance, such high counting speeds are likely to become commonplace within a short time. With this prospect in view it is felt that counting methods will provide the best method of conversion available in the near future and, further, will require the minimum of maintenance and adjustment.

## 4. Techniques for Improving the Signal-to-quantizing Noise Ratio of the Basic P.C.M. System

### 4.1 General

Many techniques for improving the signal-to-quantizing noise ratio of the basic p.c.m. system have been proposed; some of these are closely allied to the methods used to improve the signal-to-noise ratio of analogue systems, whilst others take advantage of the digital nature of the p.c.m. system. The aim of many of these techniques is to reduce the number of bits per sample used, often by taking advantage of the redundancy in the signal or by limiting the form which the signal is permitted to assume. For high-quality sound-signal transmission it is not considered worth while, at the present time, to try to exploit redundancy in the signal; further, limiting the form which the

sound signal is permitted to assume is, in general, unacceptable.

#### 4.2 *Pre- and De-emphasis*

Pre- and de-emphasis of the high audio-frequency components of the programme is commonly employed to reduce noise in the carrier circuits used for programme transmission, and in principle is applicable also to a p.c.m. system. It has been found, however, that if the degree of pre- and de-emphasis is sufficient to give a substantial reduction in noise, then the nominal maximum signal level at the input to the system will be exceeded whenever the programme contains a greater-than-average proportion of high-frequency components. A p.c.m. system, which can produce only a restricted range of digital numbers, has a more sharply defined overload level than a carrier system and can produce audible distortion if this level is only slightly or momentarily exceeded. Such overloading can be avoided by providing a quick-acting limiter<sup>12</sup> which temporarily reduces gain when required to prevent the signal level entering the a.d.c. from exceeding the prescribed maximum. If, however, a high degree of pre- and de-emphasis is applied, the protective gain reduction produced by the limiter can in itself be objectionable to the ear. Pre- and de-emphasis alone is therefore of limited value in a p.c.m. system for high-quality sound.

#### 4.3 *Companding Principle*

It is usual to specify the noise level in a sound transmission channel with reference to the maximum signal level that the channel will transmit without distortion: the level of noise that is permissible, however, is determined by its subjective effect when the signal level is low. It is therefore possible to increase the effective signal-to-noise ratio for a given channel by the use of an automatic gain control at the input of the system to bring the signal level nearer to the permitted maximum; at the receiving terminal, compensating gain changes are introduced by a second automatic gain control having characteristics complementary to those of the first. The automatic gain control at the sending end of the system reduces the amplitude range of the signals applied to it and is therefore known as a compressor; the complementary device at the receiving end of the system is referred to as an expander, and the two together as a compander.

The variation of gain in the compressor modulates the signal, producing additional frequency components; for distortionless transmission, these extra components must arrive together with the original signal components at the receiving terminal without amplitude or phase distortion, and must then be cancelled in the process of expansion.

In analogue transmission this requirement is met by syllabic companders, in which the rate-of-change gain is so slow that the additional components are confined to the frequency range which can be transmitted without serious distortion. In p.c.m. transmission, on the other hand, no such restriction is necessary provided that all additional frequency components generated by the compressor are transmitted unimpaired to the a.d.c. and, at the receiving

terminal, from the d.a.c. to the expander; there is no great difficulty in providing adequate bandwidth in the circuits concerned, and in principle it would be possible to employ other methods of companding which operate instantaneously.

#### 4.4 *Instantaneous Companding*

Instantaneous companding networks used in p.c.m. telephony circuits make use of the non-linear characteristics of diodes. To avoid waveform distortion of the received signal, the characteristics of the compressor and expander must be exactly complementary. This requirement can be sufficiently well met for telephony circuits; for high-quality programme circuits, on the other hand, the standards required are much more exacting and it has been found extremely difficult to reduce the audible distortion to tolerable proportions and to maintain this standard of quality over long periods.

Another method of instantaneous companding would be to operate the a.d.c. and d.a.c. using complementary non-linear codes; the quantizing levels could then be made closer together at low signal levels, thus reducing the quantizing noise, at the cost of wider spacing at high signal levels.

Instantaneous companding, even if carried out by a digital process, is, however, of limited value in a p.c.m. system for high-quality sound: widening the interval between quantizing levels at high signal amplitude not only increases the amount of quantizing noise but in the limit changes its character—as already noted in the case of very small signals—to something more closely resembling non-linear distortion. Experiments have been carried out in which this aspect of instantaneous companding was simulated by progressively reducing the number of digits in the pulse code as the instantaneous-amplitude of the signal increased. It was concluded that the degree of noise reduction obtainable by instantaneous digital companding without objectionable distortion being heard on critical programme material was insufficient to justify the instrumental complications which such a process would involve.

#### 4.5 *Syllabic Companding*

With syllabic companders, the gain does not vary appreciably during one cycle of the signal waveform and the reduction in the amount of quantizing noise with small signals is not obtained at the expense of increased noise or distortion at high amplitudes. It is, however, inevitable that the noise at the output of the system should vary with the signal level as the expander gain rises and falls. In telephony systems, this fluctuating noise—commonly referred to onomatopoeically as 'hush-hush-noise'—is to a large extent masked by the signal itself and even when audible can be tolerated as long as intelligibility is not impaired. However, in wide-band sound systems for the transmission of broadcast programmes, the predominant components of the signal and of the noise often lie too far apart in the a.f. spectrum for masking to be effective; moreover, even a small amount of programme-modulated noise can be aesthetically objectionable.

The subjective effects of programme-modulated noise can be mitigated by dividing the audio-frequency band into several channels by appropriate filters, each band being provided with an independent compressor and expander; the outputs from the individual compressors are combined at the sending end of the system and split between the corresponding expanders at the receiving end<sup>13</sup>. Any noise components appearing in a frequency band containing little or no signal energy are thus subject to the maximum attenuation in the expander. The practical realization of such devices, however, entails considerable instrumental complication.

Fortunately, for systems such as p.c.m., in which the noise is predominantly high-pitched, a simplified arrangement can be adopted in which only the signal components in the upper part of the audio-frequency range are subject to compression and expansion; in some cases, moreover, it is possible to achieve nearly the same result without band-splitting by introducing a large amount of high-frequency pre- and de-emphasis respectively before and after the compandor<sup>14</sup>.

#### 4.6 Influence of Compression Law

A further factor which influences the effect of programme-modulated noise in compandors is the output/input characteristic of the compressor since this determines the relationship between signal level and noise at the output of the system. If the output/input characteristic of the compressor is of the form indicated by curve (a) in Fig. 5, all incoming signals at levels above some low value  $P_1$  will

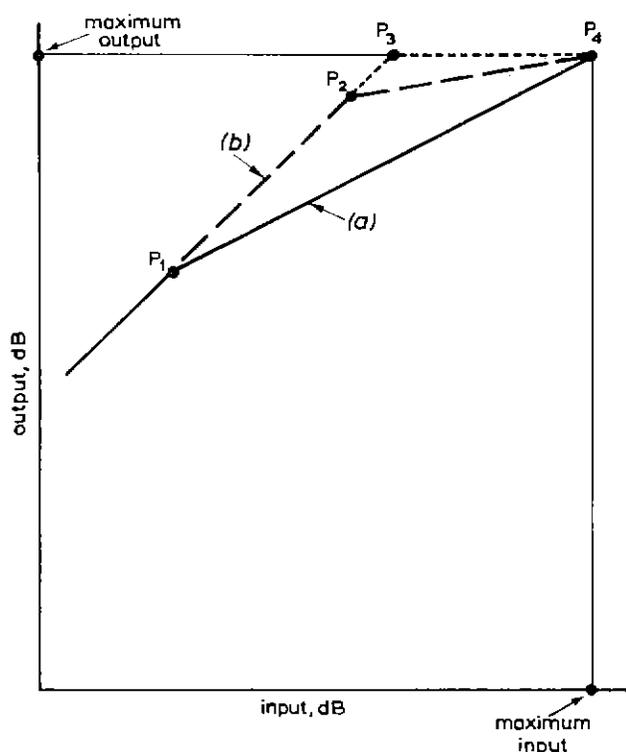


Fig. 5 — Alternative output/input characteristics of the compressor

be subject to gain reduction and their level fluctuations will produce corresponding fluctuations in noise at the output of the expander. If, on the other hand, the gain reduction occurs only at signal levels above some point  $P_2$  nearer to the permitted maximum output for the transmission system as indicated by curve (b), the resulting noise, while more sudden in its onset, will occur only during the louder passages in the programme, and is then more likely to be masked. To obtain the maximum benefit from this effect, the compression characteristic of Fig. 5 can be extended to  $P_3 P_4$ ; the compressor then performs the function of the protective limiter referred to in 4.2, although in the present case, the gain reduction is compensated by the expander at the receiving terminal. When the compressor is operating in the region  $P_2 P_4$ , the level at its output is independent of that of the incoming signal, so that some of the information necessary to reconstruct the latter at the receiving terminal is lost. The difficulty can, however, be overcome by adding to the incoming signal a pilot tone having a frequency above the audio-frequency band to be transmitted. At the receiving terminal this tone is filtered out and made to control an expander in such a way as to restore the signal level to that of the original<sup>15</sup>. An experimental p.c.m. system has been constructed using this form of compandor in conjunction with high-frequency pre- and de-emphasis as described earlier; to avoid the generation of difference frequencies through the sampling process, the pilot tone was of half the sampling frequency and phase locked. With this arrangement it was found possible to reduce the audible level of quantizing noise by up to 15 dB, equivalent to the effect of adding  $2\frac{1}{2}$  bits to the digital code.

## 5. Incorporation of the P.C.M. Sound Signal in a Video Waveform

### 5.1 General

Reference was made in Section 1 to the possibility of incorporating the sound signal associated with a television programme in the video waveform; this artifice will now be considered in more detail.

With the sound signal incorporated in the video waveform, the need for a separate high-quality sound distribution network for the 625-line television service could be avoided and the possibility of operational errors would be reduced since the sound could not become dissociated from the appropriate video signal between terminal points. Overall reliability of the sound distribution system should be improved, since failure of video links is comparatively rare and solid-state circuitry will ensure high reliability of the terminal equipment. Sound quality would be consistent and independent of the quality of the video link over wide limits and a uniform audio-frequency response would always be available.

### 5.2 Time Available in the Video Waveform for Transmission of the Sound Signal

In a monochrome signal, the total time allocated to the line and field synchronizing and blanking intervals occupies some 25 per cent of the total time required to send a

complete picture; the field blanking interval occupies 6.5 per cent of each field period, and 19 per cent of each line period is occupied by line-blanking. In a colour signal, the colour burst occupies the post-line-synchronizing blanking interval, leaving the line-synchronizing pulse as the most suitable interval for the transmission of other information; this period is therefore chosen for the p.c.m. sound signal. The line-frequency of the 625-line system is 15.625 kHz, a figure which is comparable with the upper limit of the audio-frequency range to be transmitted. It is therefore convenient to make the sampling frequency of the associated p.c.m. sound system equal to twice line frequency, using the method described later to convey two samples in each line-synchronizing pulse period.

The chief disadvantage of the line-synchronizing period is its brevity and consideration has been given to ways in which it could be temporarily extended to accommodate the sound signal. Fig. 6(a) shows the line-blanking interval of a 625-line waveform. It will be seen that, excluding the 'front porch' and colour burst, the total time available for line-synchronizing and other information is  $8.2 \mu\text{s}$ , this being made up of  $4.7 \mu\text{s}$  of line-synchronizing pulse,  $0.8 \mu\text{s}$  of black level from the trailing edge of the line-synchronizing pulse to the start of the burst, and  $2.7 \mu\text{s}$  of black level from the end of the burst to the start of the picture; the figures given are all nominal half-amplitude values. If the longest possible period for transmitting the sound signal is desired, the waveform could be modified to the form shown in Fig. 1(b), thus making available an uninterrupted  $7.8 \mu\text{s}$  period of synchronizing level (200 ns guard-space is allowed at each end of the burst). This modification was rejected since high precision would be needed to move the burst to its new position and subsequently replace it; it was considered that the process would involve a significant increase in equipment complexity, a complication justified only if absolutely neces-

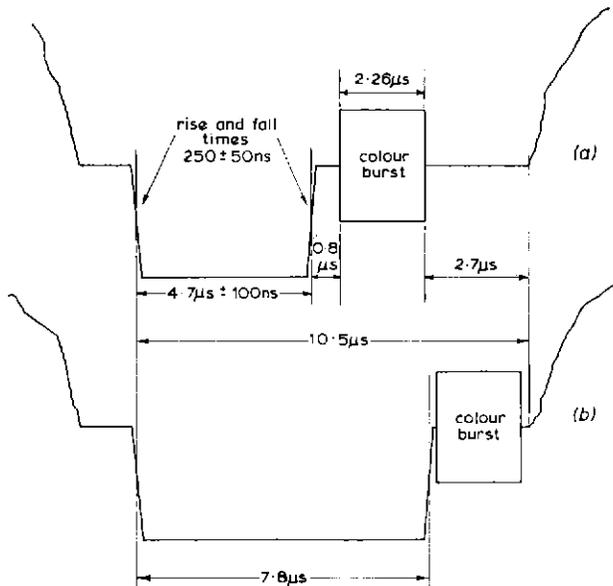


Fig. 6 — Standard and modified 625-line sync waveforms  
(a) Standard (b) Modified

sary. The conventional form of line-synchronizing waveform was, therefore, retained and the sound information inserted in the line-synchronizing intervals.

An examination of the tolerances proposed for the 625-line waveform as distributed to transmitters shows that the minimum line-synchronizing pulse duration is  $4.6 \mu\text{s}$  with maximum rise-times of 300 ns. The leading edge of the line-synchronizing pulse is the primary timing edge in the waveform. A guard band should therefore be kept between the half-height point of the leading edge and the sound signal; this ensures that the timing edge will not be disturbed by the sound signal, even when group delay distortion is present on the circuit. Consideration of the inherent delays incurred in the generation and shaping of the sound pulses, and of the need to leave some part of the bottom of the synchronizing pulse undisturbed to facilitate the process of separating synchronizing pulses from the combined signal, suggest the adoption of a 400 ns guard space. With a similar guard space preceding the trailing edge of the line synchronizing pulse, the interval remaining for the sound signal is  $3.8 \mu\text{s}$ . During the field blanking interval, alternate equalizing pulses need to be extended in the process of inserting the sound signal; at the receiving terminal, however, these equalizing pulses can be restored to the normal width of  $2.3 \mu\text{s}$ .

### 5.3 Specification of the Pulse-code Modulation System

#### 5.3.1 Sampling Frequency

As already indicated in Section 2.1, converting an analogue signal to digital form requires regular sampling at a frequency that is at least twice the highest frequency present in the original signal. It is preferable for the sampling frequency to be a simple multiple of the television line-frequency, and the choice of twice line-frequency, mentioned in the last section, gives a sampling frequency of 31.25 kHz for a 625-line system. This results in an effective audio bandwidth of 14 kHz when the imperfections of practical filters are taken into account.

#### 5.3.2 Number of Digits

The minimum number of digits likely to prove satisfactory for a practical system is 10; this corresponds to 1,024 quantizing levels. Such a p.c.m. system has a theoretical peak signal-to-r.m.s. quantizing ratio of 65 dB; however, this figure can be increased by as much as 15 dB by the use of a compandor of the type described in Section 4.6.

In order to combine groups of 10 digits that are produced at a frequency of 31.25 kHz, with a 625-line video signal, alternate groups must be stored for half the duration of a television line ( $32 \mu\text{s}$ ) so that two groups can be inserted in each line-synchronizing interval. In addition an extra pulse must be inserted immediately before each combined group of 20 digits, to act as a time reference for the decoding process.

#### 5.3.3 Digit-pulse Amplitude

The maximum pulse amplitude which can be reliably transmitted through a video circuit is equal to the excu-

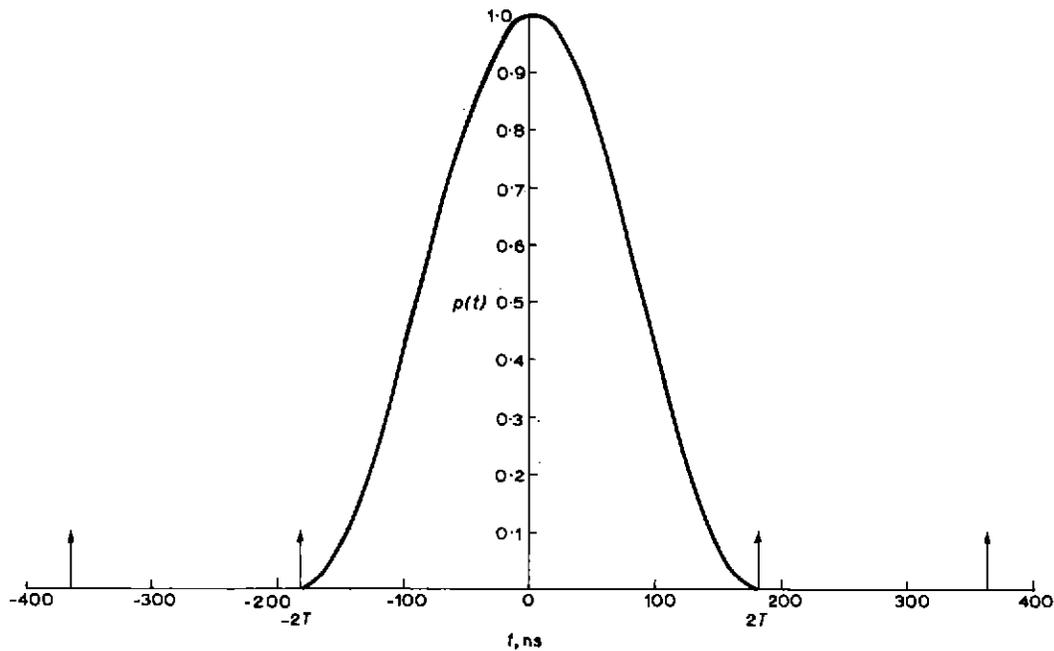


Fig. 7 — Ideal  $2T$ -cosine<sup>2</sup> pulse (bandwidth approximately 5.5 MHz)

$$p(t) = \frac{1}{2}[1 + \cos(\pi t/2T)]; \quad -2T \leq t \leq 2T; \quad 2T = 182 \text{ ns}$$

↑ Minimum spacing of pulses without reducing noise immunity

sion from synchronizing level to peak-white level. Pulses of this amplitude will provide the greatest possible protection against noise and interference and since these pulses will never be transmitted or displayed on a picture monitor there is no reason why they should not be used.

#### 5.3.4 Digit-pulse Shape and Spacing

Individual pulses of the 21-pulse combination (one marker pulse plus two 10-digit words) must each have a well-defined start and finish since any spurious disturbance outside the  $3.8 \mu\text{s}$  interval could cause interference with other parts of the waveform. An obvious choice for the pulse shape is that formed by one cycle of the cosine<sup>2</sup> function; this is the shape of the well-known  $2T$ -test pulse shown in Fig. 7. The practical pulse has insignificant minor lobes and a spectrum which falls to zero at 5.5 MHz with negligible power at higher frequencies.

The closest spacing of pulses which can be used without reducing their immunity to noise is that for which the maximum of one pulse exactly coincides with the zero of an adjacent pulse as indicated by the arrows in Fig. 7. Using this spacing, the full noise immunity can be realized only if the determination of the presence or absence of a pulse is carried out at precisely the time at which its maximum occurs. This spacing is also the closest which may be adopted if the amplitude of a block of pulses is not to be greater than the amplitude of an individual pulse, a desirable condition, since it allows pulses of the greatest possible amplitude to be used.

#### 5.3.5 Experimental Sound-in-vision System

An experimental sound-in-vision system using a 10-bit

code and incorporating the pilot tone-controlled compandor arrangement referred to in Section 4.6 has been constructed. Fig. 8 shows an oscillogram of a 625-line video waveform during the line-blanking interval, illustrating the p.c.m. signals, preceded by a timing digit, inserted in the line-synchronizing pulse. In the example, the two consecutive 10-digit samples being transmitted were identical, each having the digital code 1011000100. Interleaving the two samples, complementing odd-numbered digits (i.e. transmitting 0 in place of 1 and vice versa), and adding the timing pulse at the start of the group gives the sequence 1011001011010011010 shown.

The equipment was found to operate satisfactorily in a number of tests, using in some cases actual video links and in others, circuits simulating a link having the maximum degree of distortion permitted by its specification.

## 6. Conclusions

From a survey of known art, it is concluded that pulse-code modulation should be capable of satisfying present and foreseeable future requirements for a high-quality sound-signal distribution network.

Experimental coding and decoding equipment has been constructed to the specification given in Section 2.4. The results showed that it is possible to engineer an 11-digit system having a performance close to that predicted by theory; from the experience obtained it is concluded that an efficient 12-digit system is practicable.

It has also been shown that there is sufficient redundant time available in the 625-line video waveform to include the accompanying sound signal in the form of a 10-bit

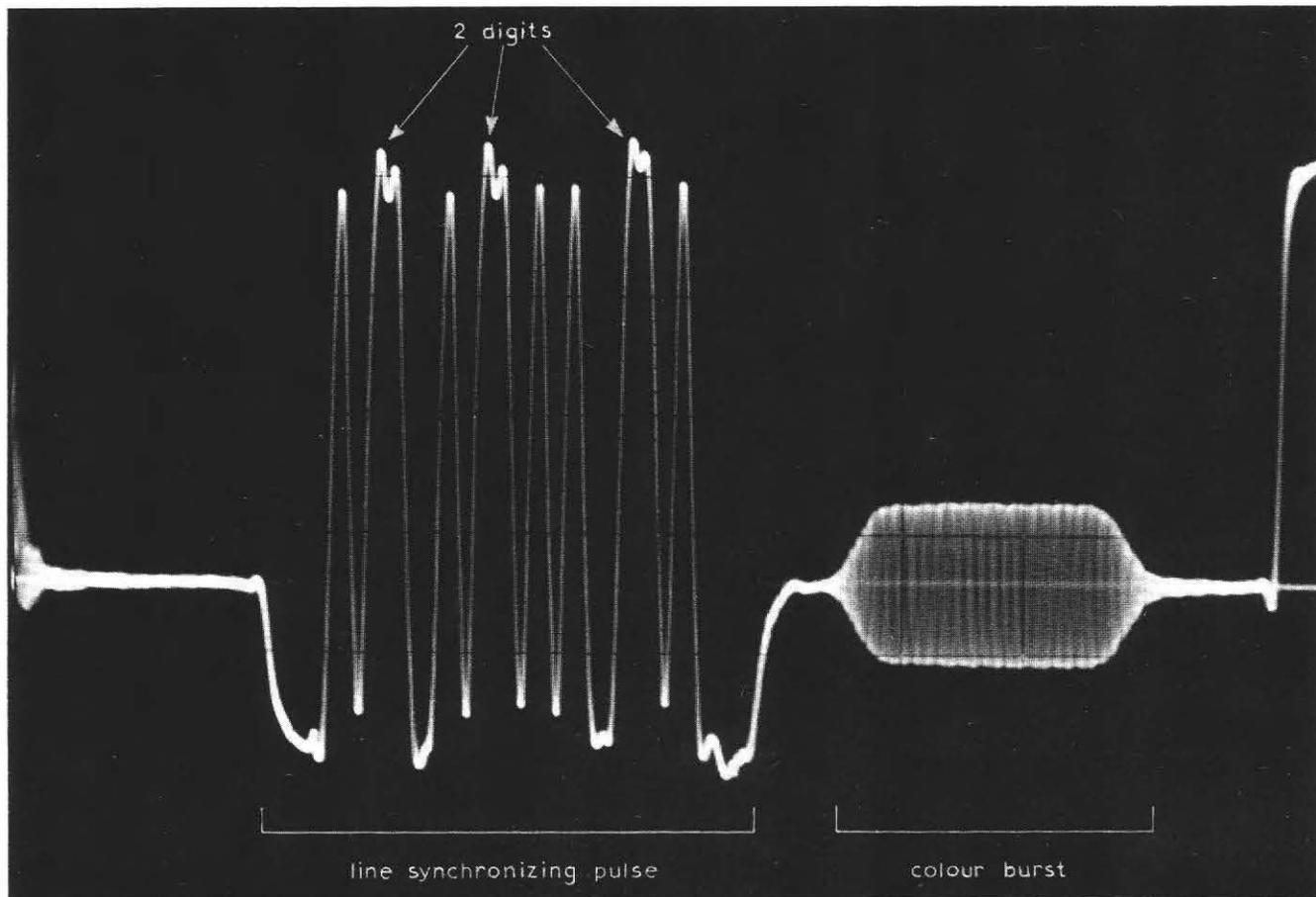


Fig. 8 — Sound-in-vision waveform showing two interleaved 10-digit samples inserted in the line synchronizing pulse period

p.c.m. signal, and that by the use of an appropriate companding system, a satisfactory sound-in-vision distribution system could be designed.

Systems operating on the lines indicated in this monograph are capable of transmitting a somewhat wider audio bandwidth than is at present provided for transmitters outside the London area, with a signal-to-noise ratio conforming to existing tolerances. The standard of performance attained is not degraded by regeneration at repeater stations, while monitoring of the links is reduced to the simple process of verifying the presence of the pulse signals.

## 7. References

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## A RECENT BBC TECHNICAL SUGGESTION

### FLARE CORRECTION USING FILTERING AND SAMPLING TECHNIQUES TO GENERATE A SIMULATED FLARE SIGNAL

The horizontal flare profile is determined by the step response of the low-pass filters 'A' and 'E', having a cut-off frequency  $f_a$ . The horizontal sampling switches operate at a frequency  $x.f_h$ , where  $f_h$  is the horizontal scanning frequency of the signal being corrected and  $x.f_h \ll 2f_a$ .

Each of the  $x$  outputs of the first horizontal sampling switch is a signal representing brightness variations down a vertical strip of the picture.

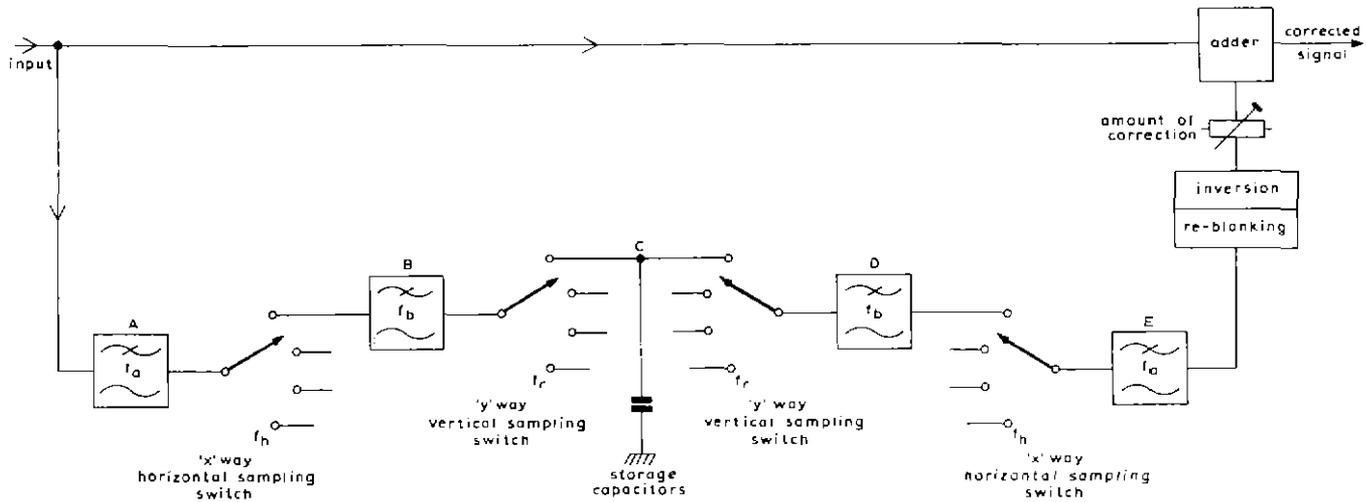
The vertical flare profile is determined by the step response of the  $x$  low-pass filters 'B' and the  $x$  low-pass filters 'D', having a cut-off frequency  $f_b$ . The vertical

sampling switches operate at a frequency  $y.f_v$ , where  $f_v$  is the vertical scanning frequency of the signal being corrected and  $y.f_v \ll 2f_b$ .

The correct line and field timing of the simulated flare signal with respect to the signal to be corrected is achieved by charge storage in the  $x.y$  capacitors 'C' and their subsequent resampling during the next vertical scanning period. The exact line and field timing is determined by the relative angular instantaneous 'positions' of the input and output, horizontal and vertical sampling switches.

The output of the simulated flare signal is re-blanked and the desired amplitude of it subtracted from the signal to be corrected in the main signal path.

B. W. PETHERS



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