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FOREWORD

"Communications and the Community of Man" was chosen as the theme of ICC-71, the seventh annual IEEE International Conference on Communications. This topic is examined in four special theme sessions which focus attention on the impact of communications technology on man and his environment.

Forty-seven technical sessions were scheduled for ICC 71. A total of 235 papers were presented, including those of members of the various panels and theme sessions. Seventy-three authors from outside Canada and the United States presented papers. These include two authors from the USSR, which marks their country's first participation in the ICC.

The Communications Technology Group is the principal sponsor of the Conference, with the Montreal Section of IEEE as co-sponsor. In addition to papers coordinated by various Technical Committees of the Communications Technology Group, other IEEE Groups, including Aerospace and Electronics Systems, Antennas and Propagation, Information Theory, Circuit Theory, Audio and Electroacoustics, participated in ICC 71.

This year, for the first time, the Conference was held outside the United States. Simultaneous translation facilities provided both English and French versions of all papers in the theme sessions. Special thanks are due to the efforts of the ICC Committee members listed in Appendix II who made the Conference a reality.

ICC 71 also featured a wide range of exhibits sponsored by major Communications Companies, which were designed to complement the technical program.

In 1972, ICC has chosen as its theme: "Communications - Versatile Tool of Mankind".

ICC-72 will be held in Philadelphia, June 2, 3, 4. Mark these dates on your calendar now!

Donald M. Atkinson
Chairman, ICC 71

ICC 71 TECHNICAL PROGRAM

All sessions will be held in the Queen Elizabeth Hotel

LOCATION	MONDAY, JUNE 14 9:00 AM-12:00 Noon	MONDAY, JUNE 14 2:00 PM-5:00 PM	TUESDAY, JUNE 15 9:00 AM-12:00 Noon	TUESDAY, JUNE 15 2:00 PM-5:00 PM	WEDNESDAY, JUNE 16 9:00 AM-12:00 Noon	WEDNESDAY, JUNE 16 2:00 PM-5:00 PM
Le Grand Salon	Session 1 **Audio and Electro-Acoustics** Delta Modulation	Session 9 **Theme Session** Socio-Economic impact of Communications Technology	Session 17 **Theme Session** Communications and the Community of Man I	Session 25 **Theme Session** Communications and the Community of Man II	Session 33 **Theme Session** Communications Technology and Education	
Marquette	Session 2 **Radio Communication** Digital Radio Communication Topics	Session 10 **Radio Communication** Spectrum Utilization and Interference	Session 18 **Radio Communication** Specialized Common Carrier Planning (Panel)	Session 26 **Radio Communication** Angle Diversity systems for scatter applications	Session 34 **Radio Communication** Propagation Problems in radio communications	Session 41 **Radio Communication** Detection Topics in Radio Communications
Richelieu	Session 3 **Circuit Theory** State of the Art in Circuit Techniques	Session 11 **Aerospace & Electronic Systems** Domestic Satelite Systems	Session 19 **Aerospace & Electronic Systems** Communications Satelite Systems Technology	Session 27 **Space Communication** Satelite Communication above 10 GHZ	Session 35 **Space Communication** Intelstat System Design and Operation	Session 42 **Space Communication** Satelite Multiple Access Methods
Saguenay	Session 4 **Communication Systems Discipline** Panel Discussion on technical control of Communication Networks	Session 12 **Communication Systems Discipline** Performance Criteria for communication systems	Session 20 **Communication Systems Discipline** Tropscatter communication systems	Session 28 **Antennas and Propagation** Antennas	Session 36 **Information Theory** Coding	Session 43 **Information Theory** Probabilistic Theory
Bersimis	Session 5 **Communication Theory** Modulation	Session 13 **Communication Theory** Detection	Session 21 **Communication Theory** Equalizers	Session 29 **Communication Theory** Channel-Characterization	Session 37 **Communication Theory** Impulse Noise Measurements	Session 44 **Data Communication** Electronic Data Switching System (EDS)
Jolliet	Session 6 **Wire Communication** New trends in analog channel banks	Session 14 **Wire Communication** Panel on the evolving digital hierarchy	Session 22 **Wire Communication** Digital Systems	Session 30 **Wire Communication** Subscriber Systems	Session 38 **Wire Communication** Transmission Media	Session 45 **Wire Communication** Panel Discussion on: Telecontrol-Telecommunications cable and interfacing Equipment in a H.V. enviroment
Mackenzie	Session 7 **Data Communication** Facsimile and line drawing systems	Session 15 **Data Communication** Advances in data communications 1971	Session 23 **Data Communication & Communication Switching** DATRAN – Data Transmission Network	Session 31 **Data Communication** Panel on communication network planning for data transmission	Session 39 **Data Communication** Data Communication Network & Systems	Session 46 **Data Communication** Digital Communication Techniques
Duluth	Session 8 (withdrawn)	Session 16 **Communication Switching** Customer Switching Systems	Session 24 **Telemetering** Modern Telemetering	Session 32 **Communication Switching** Automatic Intercept Systems	Session 40 **Communication Switching** Switching Networks	Session 47 **Communication Switching** Switching and Measurement Systems
Bersimis		Session 48 Monday evening Caucus 7:30 PM-9:30 PM **Data Communications** Display Function Control Codes				

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QUANTIZING NOISE CALCULATIONS IN DELTA MODULATION*

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ABSTRACT

A new approach for determining quantizing noise power in delta modulation has been developed. This approach has been applied to the delta modulation of Gaussian first order Markov sources. The analysis procedure avoids the traditional and somewhat artificial technique of dichotomizing the quantizing noise into slope overload and granular components, and then working with these two components independently. Formulae, which can be evaluated numerically, are developed for quantizing noise probability density functions and power spectra. Computer solutions are compared with previous simulation results. Although the procedure described herein produces accurate calculations for quantizing noise at all values of step size, it is computationally difficult to apply. This paper is a condensation of research reported in reference [4].

INTRODUCTION

Delta Modulation (ΔM) is perhaps the simplest and most economical way to encode analog signals or information into digital form for digital transmission or processing. Source encoding systems like ΔM, pulse code modulation (PCM), differential PCM (DPCM), etc., always introduce a degradation called quantizing noise. This paper presents a new method for calculating the quantizing noise power for ΔM systems with random input signals. Such calculations are difficult because ΔM is a non-linear feedback system.

Previous calculations of quantizing noise power in ΔM have been based on the assumption that the quantizing noise can be split into two statistically independent components--granular noise and slope overload noise [1,2]. This approach assumes that quantizing noise is due to its granular component most of the time. When slope overload noise occurs, it occurs only in short bursts. During a burst, the overload noise power is so large that the granular component is considered to be negligible. The assumption that these two noise components are independent allows them to be treated separately and their variances added to find total quantizing noise power. Signal-to-quantizing noise ratio calculations based on this technique are valid for many cases of practical interest. However, theoretical results do not agree very well with computer simulation for small quantizer step sizes. It is under this condition that quantizing noise is predominately due to the slope overload component.

The value in the new approach presented here lies in the fact that this somewhat artifical dichotomy of the quantizing noise is avoided for the important case of Markov-Gaussian information sources. This case is important because ΔM systems are ideally suited for encoding sources of this type. Furthermore, a Markov process is a reasonably good model for a television signal, a prime candidate for ΔM encoding.

*This work was supported by the Air Force Office of Scientific Research, Office of Aerospace Research, United States Air Force, under contract F44620-69-C-0033.

The novelty of this noise analysis scheme is that the information source is allowed to "start up" from a condition of inactivity. Because of this, an exhaustive sequence of possible quantizing noise probability density functions corresponding to an arbitrary output bit stream (from the delta modulator) can be generated with recursion formulae. Transients associated with "start up" decay quickly and the statistical properties of the quantizing noise stabilize after only a few samples are fed into the ΔM. Suitable averaging over the generated ensemble produces an overall quantizing noise density. This same general procedure can be used to compute an ensemble of possible conditional densities which ultimately leads to calculation of quantizing noise power spectra. This technique can be generalized to apply to differential PCM as well.

The strength of the new technique is that it allows the quantizing noise power to be computed with as much accuracy as desired for any step size. Its weakness stems from the fact that it is an iterative technique which, for high computation accuracy, may require considerable computation time. This is principally due to the large number of convolution operations required. The problem is compounded in multi-level differential PCM.

In what follows the new approach for computing ΔM quantizing noise power is explained for Gaussian Markov signals and results are compared with previous computations and computer simulation results.

THE GAUSSIAN MARKOV MESSAGE PROCESS

Figure 1 presents a model of the message process to be considered. This model shows how the sequence of signal sample values $\{S_i\}$ can be created from a sequence of independent zero mean Gaussian random variables $\{x_i\}$. The relationship between the two sequences is given by

$$S_i = \alpha S_{i-1} + x_i, \qquad (1)$$

where α is a positive constant less than 1. The constant α characterizes the message source. The autocorrelation of the sequence $\{S_i\}$ is

$$E\{S_i\ S_j\} = \frac{\alpha^{|i-j|}}{1-\alpha^2}\sigma_x^2 \qquad (2)$$

where σ_x^2 is $E\{x_i^2\}$.

The sequence $\{S_i\}$ is statistically identical to sample values of RC filtered Gaussian noise. For if flat Gaussian noise is passed through a low-pass RC filter (first order Butterworth filter) its power spectrum (single sided) is

$$G(f) = \frac{(\frac{2}{\pi f_c})}{1+(\frac{f}{f_c})^2} \qquad (3)$$

where f_c is the corner frequency. The numerator has been choosen to make the total power equal to one watt. The Fourier transform of (3) gives the autocorrelation of the process

$$R(\tau) = \exp\{-2\pi f_c|\tau|\}.$$

If this signal is sampled at a rate of f_s then the autocorrelation $R(\tau)$ of the sample values will be defined only at $\tau = \frac{k}{f_s}$; $k = 0, \pm 1, \pm 2, \ldots$. Making this substitution in $R(\tau)$ above gives

$$R(k) = \exp\{-\frac{2\pi f_c}{f_s}|k|\}. \qquad (4)$$

Comparing the autocorrelation of the discrete model of Fig. 1 given by (2) with that of the sampled RC signal given by (4) shows that if $\sigma_x^2 = 1-\alpha^2$ and $\alpha = \exp(-\frac{2\pi f_c}{f_s})$ the two autocorrelation functions are equal. Since the sequences are Gaussian in both cases, identical autocorrelation implies that all statistics of the two processes are identical.

The spectrum of (3) can be used to describe the envelope of the video part of the power spectrum of a television signal. An actual video power spectrum would be determined by the picture material and would have complicated line structures at the line and field rates but (3) is a good model for its envelope. For NTSC television the value of f_c is in the neighborhood of 50 kHz.

DELTA MODULATOR MODEL

The Delta Modulator modeled in Fig. 2 is the source encoder to be considered. Members of the sequence $\{S_i\}$ are produced by the Gaussian-Markov information source of Fig. 1 and presented successively as inputs to the Delta Modulator of Fig. 2. Corresponding to a given member of the input stream, say S_i, a linear prediction of S_i, $\hat{S}_i$, is made and subtracted to form a difference random variable. The difference random variable, ε_i, is quantized to $+\delta$ $(-\delta)$ if ε_i is greater (less) than zero. Quantizing noise is, of course, introduced in this operation and so the quantizer output is shown as $\varepsilon_i + q_i$. The random variable q_i is of chief concern in this paper. Methods will be developed for calculating the probability density function (pdf) of the quantizing noise q_i. It is assumed at the outset that random variables in the sequence $\{q_i\}$ are stationary and have identical statistical properties if i is large.

Examination of Fig. 2 shows that the prediction $\hat{S}_i$ is $\hat{S}_i = \alpha(S_{i-1} + q_{i-1})$, and the input to the quantizer can be expressed as

$$\varepsilon_i = S_i - \alpha\,(S_{i-1} + q_{i-1}). \qquad (5)$$

Since the information source of Fig. 1 is being used, Equation (5) can be simplified to (see Equation (1))

$$\varepsilon_i = x_i - \alpha q_{i-1}. \qquad (6)$$

since x_i and q_{i-1} are independent, the pdf of ε_i can be calculated by the convolution of the pdf's of x_i and αq_{i-1}[3]:

$$p_{\varepsilon_i}(x) = \frac{1}{\alpha}\, p_x(x) * p_{q_{i-1}}(\frac{x}{-\alpha}) \qquad (7)$$

where $p_{\varepsilon_i}(x)$ = pdf of ε_i

$p_x(x)$ = Gaussian pdf of x_i ; zero mean, variance σ_x^2

$p_{q_{i-1}}(x)$ = pdf of q_{i-1}

and * denotes convolution. The importance of (7) to what follows cannot be over emphasized. It relates the present input pdf p_{ε_i} to the previous quantizing noise pdf $p_{q_{i-1}}(x)$. As the quantizing noise pdf can be obtained from Equation (7) once the quantizer step is specified, a recursion formula can be found which leads to noise probability density functions.

In order to develop such a recursion formula, an initial quantizing noise pdf must be assumed. This can be accomplished by allowing the information source of Fig. 1 to "start up" from a condition of inactivity. Specifically, the source will be required to "start up" at a time denoted by $i = 1$; i.e., x_i is zero for $i \leq 0$ and x_i is a zero mean Gaussian random variable with variance σ_x^2 for $i > 0$. As a consequence of "start up" $S_o = \hat{S}_1 = 0$, $q_o = 0$, $\varepsilon_1 = x_1$ and the quantizer input density at $i = 1$, $p_{\varepsilon_1}(x)$, is zero mean Gaussian with variance σ_x^2. The quantizing noise pdf at $i = 1$, $p_{q_1}(x)$ is easily determined to be one of two truncated, shifted versions of $p_{\varepsilon_1}(x)$; which one depends on the quantizer output step. This is illustrated in Fig. 3. The quantizer input pdf, $p_{\varepsilon_1}(x)$, is shown in Fig. 3a, while Fig. 3b and Fig. 3c give the resultant quantizing noise density if the quantizer exercises a step of $+\delta$ or $-\delta$, respectively. (In either eventuality, Equation (7) can be used to find the next quantizer input density.) Next, $p_{\varepsilon_2}(x)$ can be used to find $p_{q_2}(x)$ using the technique of Fig. 3 if a second output step is assumed.

Once a sequence of quantizer output steps is assumed, future quantizing noise pdf's can be found by recursion formulae to be developed as outlined above. Allowing all output sequences of length n will result in an ensemble of all possible quantizing noise pdf's that can occur n units of time after "start up". By suitable averaging, pdf's for $q_1, \ldots, q_n$ are obtained. If the information source and Delta Modulator have been operating long enough (large n), starting transients will have died out and the pdf for q_n will also characterize q_{n+k} $k = 1, 2, \ldots$. I.E.,

$$\lim_{i \to \infty} p_{q_i}(x) = p_q(x)$$

where $p_q(x)$ is the true probability density of the quantizing noise.

THE NOISE TREE

The ensemble of possible quantizing pdf's corresponding to Delta Modulator operation for n units of time after "start up" can be organized into a graph that will be called a Noise Tree. A Noise Tree, as shown in Fig. 4, is defined by parameters α, δ, σ^2. This structure has two component parts: nodes and branches. Nodes are labeled by a double subscript notation as N_{ij} where i denotes its column and j denotes its position from the top of a column. Branches connect nodes in adjacent columns as shown. Passage of time, from start up, occurs left to right beginning at node N_{11}. Operation can proceed from N_{11} to either N_{12} or N_{22} by the connective branches if the quantizer selects a $+\delta$ or $-\delta$ respectively. As a similar choice can be then made at either N_{21} or N_{22}, the Delta Modulator can be at any four nodes (N_{31}, N_{32}, N_{33}, N_{34}) after two units of time. Obviously there is one path, consisting of nodes and branches, through the tree for every string of n output symbols from the Delta Modulator. Hence, calculation of the performance at each node will be exhaustive for output strings of length n.

At each node in the Noise Tree, two probability densities are defined--the quantizing noise pdf and the quantizer input pdf. And, at each node, these pdf's can be computed from the quantizing noise pdf at the previous node. By assuming a known quantizing noise pdf at N_{11} the quantizing noise pdf can be calculated at all other nodes in the tree. Since the system starts up from a condition of inactivity, the quantizing noise at node N_{11} is assumed to be zero, i.e., $q_o = 0$.

Recursion formulae based on (7) can be developed from which the quantizing noise power spectral density, and therefore the quantizing noise power can be computed. The development of these recursion formulae and computational details are too complicated and lengthy to be presented here. They are the subject of reference [4].

The next section presents results obtained by using this iterative technique to compute quantizing noise probability density functions, power spectra and signal-to-quantizing noise ratios for some ΔM systems of interest.

RESULTS

Quantizing noise pdf's were computed by constructing a number of Noise Trees using the procedure of Section IV where the step size, δ, was made variable and other parameters were held constant. Fig. 5 shows a family of quantizing noise pdf's obtained for different quantizer step sizes. In this case the feedback gain, α, was fixed to reflect sampling a Gaussian-Markov source at a rate of 50 times its corner frequency, i.e., $f_s = 50\ f_c$. Signal input to the delta modulator was normalized so that $\sigma_x^2 = 1$. The Noise Trees were generated to a length of n = 11 (Fig. 4) and $p_{q_i}(x)$ was computed for i = 1, ... , 10. In all cases, first and second moments of $p_{q_9}(x)$ and $p_{q_{10}}(x)$ agreed to at least three decimal places. The Noise Trees were not extended further because densities for smaller values of i were approaching $p_{q_9}(x)$ and $p_{q_{10}}(x)$ in shape and these two densities were indistinguishable when drawn on the same set of axes. Starting transients, therefore, were assumed to have decayed by this time and $p_{q_{10}}(x)$ was taken to be $p_{q_i}(x)$ for i>10 denoted simply $p_q(x)$. Minimum mean square error (variance of $p_q(x)$) occurred when the quantizer step size $\delta = 0.5$. This agrees with previous computer simulation results (see Fig. 5, curve for $F_s = 2$, ref. [1]).

Quantizing noise power spectra were calculated according to the method explained in reference [4]. This involves using conditional pdf's to develop a conditional noise tree similar to the noise tree of Fig. 4. Autocorrelation functions and power spectra can be computed using this technique.

Typical of power spectra derived by this approach is that of Fig. 6. The case shown is for a sampling rate f_s which is 100 times the corner frequency f_c. The signal whose spectrum is given by (3) is being sampled at a point where its power is approximately 20 log 100 or 40 dB down from its maximum value. Fig. 6 shows that if the step size δ is large ($\delta = 2.0$), the quantizing noise tends to become large at $\frac{1}{2} f_s$. The ΔM tends to oscillate transmitting long sequences of $+\delta$, $-\delta$, $+\delta$, $-\delta$, ... , etc. At some intermediate value of step size $\delta = 0.5$ in this case, the quantizing noise is relatively flat. As the step size gets smaller ($\delta = 0.125$) the quantizing noise spectrum begins to approach the spectrum of the signal itself and, like the signal, most of its power becomes concentrated in the low frequencies.

In most previous studies of ΔM the signal is bandlimited to some cutoff frequency f_o and them sampled at a rate greater than $2f_o$. The model of the Gaussian Markov process discussed in Section II does not produce sample values which come from a precisely bandlimited process (such processes are but a useful fiction, anyhow). In order to obtain meaningful quantizing noise power calculations for comparison with previous results, a bandlimiting frequency of $f_o = 25\ f_c$ will be assumed (any value of f_o greater than about 10 f_c would be satisfactory). Quantizing noise occurring at frequencies greater than f_o will be assumed to be out of band. Only that quantizing noise in the frequency band (0, f_o) will be in band. Fig. 7 shows a plot of signal-to-quantizing noise ratios S/N versus normalized step size $\delta \frac{f_s}{f_o}$ for several values of f_s where f_o is assumed to be 25 f_c. These results are compared with previous analytical and computer simulation results reported in [1]. As shown in Fig. 7 the new calculations predict the computer simulation results much more faithfully than previous analytical results when the step size is small, i.e., in the region of slope overload noise. Unlike previous approaches the one reported here does not artifically dichotomize the noise into two components and work with each separately. Good results are obtained for all values of step size.

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[3] A. Papoulis, Probability, Random Variables, and Stochastic Processes, McGraw-Hill Inc., New York, New York, 1965.

[4] T. K. Mills, "A Different Approach to Quantizing Noise Analysis in Delta Modulators," Project THEMIS Report, Electrical Engineering Department, N. C. State University, Raleigh, North Carolina, June 22, 1970.

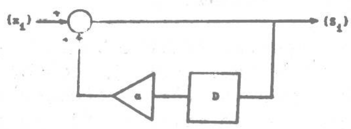

Figure 1. Model of the Gaussian Markov Information Source. D is a unit delay and a is the gain constant, 0<a<1.

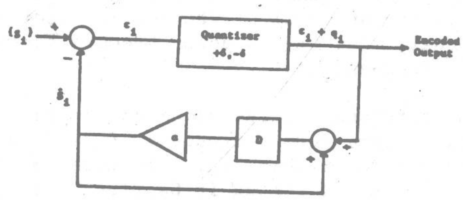

Figure 2. Delta Modulator Model to be considered (D is a unit delay and a is the gain constant, 0<a<1)

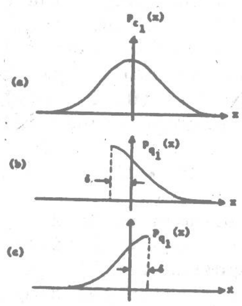

Figure 3. Probability density functions associated with "start up"
(a) Quantizer input density at i = 1
(b) Quantizing noise density at i = 1 when positive step (+δ) is exercised
(c) Quantizing noise density at i = 1 when negative step (−δ) is exercised

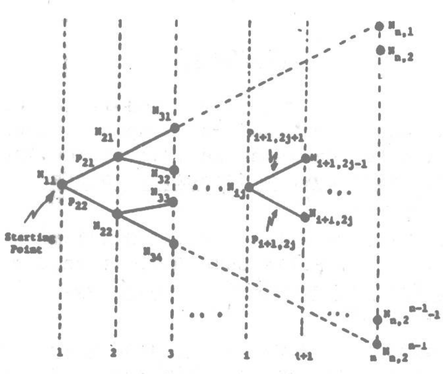

Figure 4. Noise Tree of length n. Delta Modulator "starts up" at node N_{11} and proceeds to either N_{21} or N_{22} if the quantizer transmits +δ or −δ respectively.

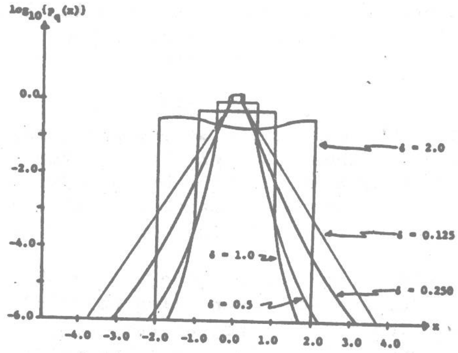

Figure 5. $\mathrm{Log}_{10}(p_q(x))$ as a function of x for δ = 0.125, 0.25, 0.5, 1.0 and 2.0. Sampling rate f_s is 30 times the corner frequency f_c.

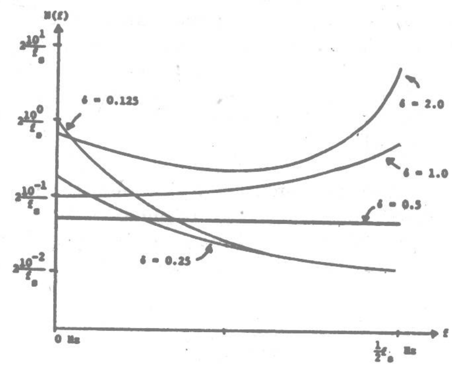

Figure 6. Quantizing noise power spectra for step sizes; δ = 0.125, 0.25, 0.5, 1.0 and 2.0, f_s = 100 f_c

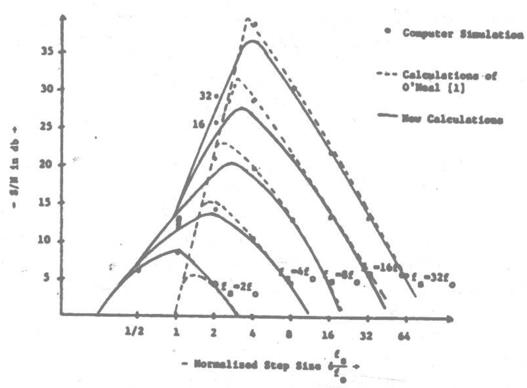

Figure 7. Signal-to-noise ratio as a function of normalized step size for sampling rates f_s = 2, 4, 8, 16 and 32 f_o. f_o is assumed to be 25 f_c.

PERCEPTUAL EFFECTS OF CODE MODULATION FOR ACOUSTICAL SIGNALS

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ABSTRACT

Perceptual effects of code modulation systems (PCM and DM) for the acoustical signals (speech and music) are discussed for the following three coding systems, uniform coding, instantaneous companded coding or non-linear coding and envelope-companded coding. Improved performance can be obtained by matching the coding system to the spectrum of the acoustical input signal. The results of transmission quality tests for voice transmission are given for systems limited in the amplitude and for systems limited in slope of this signal.

In these tests the input signal of the coding system is varied from -10 dB to +20 dB which means 10 times overloading in amplitude or slope of the signal.

A comparison is made between PCM and DM with respect to the perceptual effects of digital errors caused by noise and interference in the transmission path.

Transmission quality scores for speech transmission are given for PCM and DM with digitally envelope-companding utilizing the bit rate of 40 kb/s and for the bit rate of 24 kb/s.

High-quality music transmission is obtained for a bit rate of 256 kb/s.

INTRODUCTION

Code modulation for the transmission of acoustical signals in digital form started about 25 years ago with uniform coding for speech signals. Two well known basic systems are used, viz. pulse-code-modulation (PCM)[1,2,3,4,5] and delta-modulation (DM)[6,7,8,9,10]. As the information is transmitted in digital form, it is inevitable that the reconstructed analogue signal will include a quantization distortion component. The problem of code modulation is to keep the quantization distortion within acceptable limits utilizing the lowest possible bit rate for transmission. The two principal measures available to achieve this aim are spectral matching and companding.

The above measures are applicable to both PCM and DM and their simultaneous applications to a system give optimum results.

Spectral matching of a coding system to the spectrum of the input signal of the encoder started with differential pulse code modulation (Diff. PCM)[11,12,13] and with delta modulation with double integration[14,15].

Companding is used to obtain a more constant S/N ratio over a large range of input signals. For uniform coding the S/N ratio will vary linearly with the level of the input signal because the quantizing noise is only dependent on the size of the (constant) quantizing steps and thus will be constant at the output of the encoder. Two methods of companding offer themselves i.e.: instantaneous companding and envelope companding.

Examples of instantaneous companding are logarithmically non-linear PCM coding systems[16,17,18] and non-linear DM coding systems[19,20]. Non-linear PCM coding is a coding principle in which the magnitude of the quantizing levels is adapted to the instantaneous value of the input signal. Non-linear DM coding is a coding principle in which the magnitude of the quantizing steps is adapted to the difference in value between two or more successive samples of the input signal. For both non-linear systems PCM and DM, large values of the input signal are encoded with relatively larger values for the quantizing steps q than in a linear encoder, whereas for small values of the input signal smaller values for q are applied. The total number of quantizing steps is determined by the bit rate utilized by the coding system. Piece wise linear coding using 13 segments results in a 24 dB higher S/N ratio at low signal levels at the expense of a 10 dB lower S/N ratio at higher input levels as compared with a uniform coding system utilizing the same bit rate[21].

Envelope-companded coding systems are coding-systems with variable magnitude of the quantizing steps, controlled according to the mean level of the analogue signal (e.g. the envelope) during a certain averaging period. For speech transmission this period is 3-5 ms[22] and for music this period is of the order of 1 ms.

If this averaging period is chosen to be equal to the typical duration of a syllabe, the system is called syllabically companded[23,24].

It is required that the expander characteristic compensates the compressor characteristic precisely. An advantage of envelope companding is that tracking errors will cause level variations only, while tracking errors in instantaneous companding give rise to non-linear distortion. However, certain stability problems may occur in envelope companding if high compression rates are used over a wide dynamic range. The best results are obtained by digitally controlled companding as described below.

DIGITALLY CONTROLLED COMPANDING

Fig. 1 illustrates the principle of operation. The encoders on this block diagram can depict either PCM or DM encoders. The analogue input signals x is encoded in the linear encoder. The magnitude of the quantizing steps applied to this encoder is varried from the maximum value q_m to a minimum value 0.01 q_m. This variation of q is controlled by the modulation level analyser circuit, which changes q linearly with the average number of sampling periods over which the modulation level of the encoder exceeds a certain amount of its maximum value, e.g. "one half". The transmitted and re-generated pulses y' are decoded in a linear decoder. The quantizing steps applied to this decoder are controlled by an identical modulation level analyser circuit connected to the re-generated bit stream.

This arrangement facilitates very high tracking accuracy and therefore high compression rates because the same digital information is employed to control the compression and the expansion action.

The feedback control at the transmitter, the forward control at the receiver and the modulation level analyser circuit at both terminals using the identical bit stream ensure an excellent response of the companding system. This is true as long as the encoder is linear and is not overloaded and therefore the modulation level analyser reacts on half the maximum modulation level of the encoder [25].

In order to compare the performance of the two different companding methods the companding characteristic of both methods are plotted in fig. 2. relative to the magnitude of the quantizing steps applied for uniform coding. Curve a shows the modulation level as a function of the input level for digitally controlled companding ($q_{min} = 0.01\ q_{max}$). Curve b gives the same characteristic for instantaneous companding (piece-wise-linear PCM 13 segments). Curve c for uniform coding is included for comparison.

THE INFLUENCE OF SPECTRAL MATCHING

Spectral matching refers to matching between the overload characteristic of a code modulation system and the spectrum of the acoustical signal. In the following, the importance of spectral matching will be illustrated with examples of code modulation systems intended for speech transmission.

The spectrum of speech signals over telephone links covers the frequency band of 200-3600 Hz and falls at the rate of approximately 6 dB per octave above about 500 Hz. For amplitude-limited coding systems i.e. its characteristic of overload as a function of frequency is flat, the quality of speech transmission is lower than for slope-limited coding systems. Slope-limited coding systems means the overload characteristic falls 6 dB/octave within the frequency range of 200 to 4000 Hz. This improvement is obtained if the coding system has quantization or/and is overloaded. Fig. 3 gives the quality score for speech transmission. The mean score of 16 listeners is given for the reference qualities, 5 for excellent, 4 for good, 3 for fair, 2 for poor, 1 for bad and 0 for non-intelligible.

The following four systems are tested.

a a slope-limited code modulation system with S/N ratio > 50 dB

b an amplitude-limited code modulation system with S/N ratio > 50 dB

c the same as a but with 20 dB S/N ratio

d the same as b but with 20 dB S/N ratio.

The input levels are -10 dB, 0 dB, +10 dB and +20 dB. Zero dB means the level where saturation just starts. Given is the mean value of the scores as a function of the input level of the system.

Matching to the spectrum of a speech input signal is achieved by an integrating network in the feedback loop of the encoder having an attenuation characteristic as shown in fig. 4, curve a. Curve b is the attenuation characteristic for the network used for music transmission. Because of the double integration action of this network within the transmitted audio band there is an extra memory action which makes it possible to approximate the analogue signal in a better way i.e. over several sampling period. Therefore the quantizing noise for these systems is not only lower as compared to systems without integrating networks in the feedback loop, but the quantizing noise also shows a less pronounced correlation with respect to the input signal and to the sampling frequency. For PCM, only single integration can be used. The measured gain in S/N ratio will be about 6 dB by this equalisation networks as compared to a system with a flat overload characteristic.

THE INFLUENCE OF COMPANDING

Beside the influence of the quantizing noise one of the most important effects in coding speech and music signals is the existence of a threshold for small values of the input signal. This threshold is caused by the value of the first quantizing level of the encoder.

A threshold level of 1% of the maximum signal level gives already a decrease in quality of speech transmission and for a threshold level of 10% the quality is so bad that the system cannot be used for speech transmission.

A simple method of reducing this influence is adding white noise or a signal with a frequency outside the transmitted low-frequency band to the input signal. The level of this added signal has to be equal to one quantizing step of the encoder.

A corresponding effect is obtained by "centre clipping". Centre clipping in code modulation means that the encoder DC input level is so adjusted that each positive value of the input signal gives a positive quantizing step and each negative value a negative quantizing step. Thus for very low input magnitudes the decoder delivers a fully clipped signal with zero crossings quantized in time. The intelligibility of clipped speech signals is much higher than for a speech signal which has passed a threshold voltage. These methods are however non-linear for small input magnitudes and to keep a system adjusted to centre clipping is a very difficult matter.

By companding, the quantizing steps for small input signals are made 100 times smaller than for uniform coding and this gives a large improvement in speech transmission quality.

This improvement is equal for PCM and DM, if identical companding characteristics are used. The utilized bit rate of the coding system has no influence on the companding characteristic as long as this bit rate is sufficiently high for the companding action.

If a uniform coding system and an envelope companding system both have the same S/N ratio for a sine wave as an input signal (fully loaded), the score for speech transmission quality tests is much higher for the companding system than for the uniform coding system. This difference in quality is caused by the difference in influence between a constant noise level and a kind of modulation noise on speech and music signals.

The difference in quality of speech transmission between uniform coding and digitally envelope companded coding with a level analyser which reacts on half the maximum modulation level is iluustrated by fig. 5.

This figure shows the quality score for speech transmission for an analogue transmission system using digitally envelope companding. The analogue signals is first digitalized in an envelope companded DM encoder using a sample frequency of 200 kHz and decoded in a uniform DM decoder to obtain a digitally envelope-compressed analogue signal. As uniform DM decoder a long-pass filter (0-4000 Hz) is used.

The quantizing noise at the output of this filter is $>$ -60 dB. This compressed signal has the same compression characteristic as shown in fig. 2a. To this compressed speech signal white noise is added and at the receiving end both signals are multiplied by the control signal of the compressor in an analogue expander to simulate a companded coding system with variable quantizing noise. Transmission quality tests are carried out by 16 listeners. The mean score for the same quality references as used for fig. 3 are plotted in fig. 5 for the S/N ratios in the transmitting path of 6 dB, 9 dB, 12 dB, 15 dB and without noise. Curve a holds for a slope-limited system with 40 dB companding and curve b holds for amplitude-limited uniform coding systems.

The quality remains fair for companded coding system with slope limiting, even for low S/N ratios. This means that the quality of speech transmission of coding systems remains acceptable for low S/N ratios if this noise resembles white. For coding systems utilizing a low bit rate however the quality of the speech transmission is determined by the threshold effect of the encoder.

COMPARISON OF PULSE CODE MODULATION AND DELTA MODULATION

As shown spectral matching and envelope companding for PCM and DM have the same influence on the quality of speech transmission for both systems.

For voice transmission DM gives a better quality than PCM when a bit rate is used $<$ 40 kb/s.

For bit rates $<$ 24 kb/s non-linear coding is nearly impossible if a high companding rate is desired. The results for spectrally matched coding systems with digital envelope companding are shown in fig. 6 for the bit rates of 40 kb/s and 24 kb/s. The systems are tested using four codecs in series in accordance with document XII/116 of the C.C.I.T.T. for three input levels, viz. 0 dB, -7 dB and -14 dB.

Twenty test persons were asked whether the system was satisfying (i.e. excellent, good or fair) or not.

This resulted in a percentage of satisfied persons for each test. Because of the limited number of test persons this estimate is accompanied by a confidence interval. This interval, indicated by bars in fig. 6, represents the resulting percentages of satisfied listeners with a confidence level of 95%.

Tested are the following systems.

Ref.syst. stands for the Reference ARAEN Filter defined by document XII/116.

a DCPCM 40 kb/s Digitally envelope companded PCM using the bit rate of 40 kb/s
b DCDM 40 kb/s Digitally envelope companded DM using the bit rate of 40 kb/s
c DCPCM 24 kb/s as a but using 24 kb/s
d DCDM 24 kb/s as b but using 24 kb/s.

For PCM the bandwidth of the transmitted analogue signal is limited by the sampling frequency of the coding system. For DM however this bandwidth is limited by the S/N performances of the used sampling frequency. So for a bit rate of 64 kb/s high-quality speech transmission is possible with a S/N ratio of 40 dB at a bandwidth of 6 kHz for the audio signal using DM. This bandwidth is 4 kHz for PCM.

Music transmission utilizing a bit rate of 256 kb/s is obtained with PCM coding having an envelope companding of 26 dB (spectral matching).

The S/N ratio is about 50 dB. The idle noise is -70 dB and the quality is high with the bandwidth of 15 kHz. Utilizing the same bit rate a similar DM coding system gives a high quality of music transmission for a bandwidth of 20 kHz.

THE INFLUENCE OF DIGITAL ERRORS

A digital error caused by noise and interference in the digital transmission path of a PCM system causes a spike in the analogue output signal of the decoder if one of the most significant bits is an error. These spikes are equal for uniform or non-linear coding systems and lower in amplitude for envelope-companded systems. For DM these errors cause a change in quantizing noise dependent on the error probability. For envelope-companded DM the influence of the errors is reduced by the same amount as the quantizing noise is reduced by the compander. So the influence of digital errors in the transmission path on the quality of speech transmission is negligible for error probabilities of less than 10^{-3}.

This effect is illustrated in fig. 7. showing the results of logatom intelligibility tests for a DM system with spectral matching and envelope companding called DCDM using a bit rate of 20 kb/s. Thus a transmission bandwidth of 10 kHz must be used for polar binary transmission.

The intelligibility score in percentages versus input level is given. The error probability is taken as the parameter. Ten procent corresponds to a S/N ratio in a 10 kHz transmission band of 2 dB, 5% to 5 dB and 1% to 7,5 dB. "Ideal" corresponds to no noise. The results show that still intelligible speech transmission is possible for a S/N ratio of 2 dB in the transmission path over a dynamic range of the input signal of 30 dB. This is not possible with standard FM voice channels.

CONCLUSION

The performance of code modulation systems can be considerably improved by means of spectral matching and companding. A system well matched to the spectrum of voice or music signals will give approximately a 6 dB higher S/N ratio than one matched to a flat spectrum.

Envelope companding is superior to instantaneous companding because the quantizing noise is less correlated to the analogue signal and to the sampling frequency. This is important for the transmission quality of speech and music signals.

The only way to determine the quality of transmission for acoustical signals is to use perceptive quality tests. This is especially true for digital errors in the transmission path and for saturation of the coding system.

Standard telephone quality voice transmission is possible at a bit rate of 40 kb/s. Intelligible voice channel transmission is possible at a bit rate of 16 kb/s. For both bit rates the system may be affected by a high noise level on the transmission path. High quality music transmission is obtained with a bit rate of 256 kb/s and a low frequency bandwidth of 15 kHz for the sound channel with a coding system (PCM or DM) matched to the spectrum of the input signal and to the level variations of this input signal.

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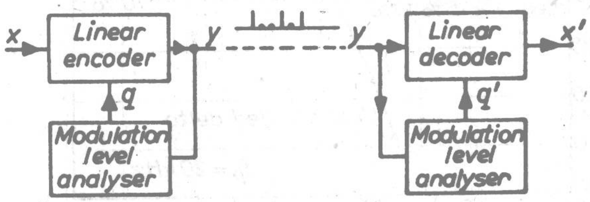

FIG.1 Principe of digitally controlled companding

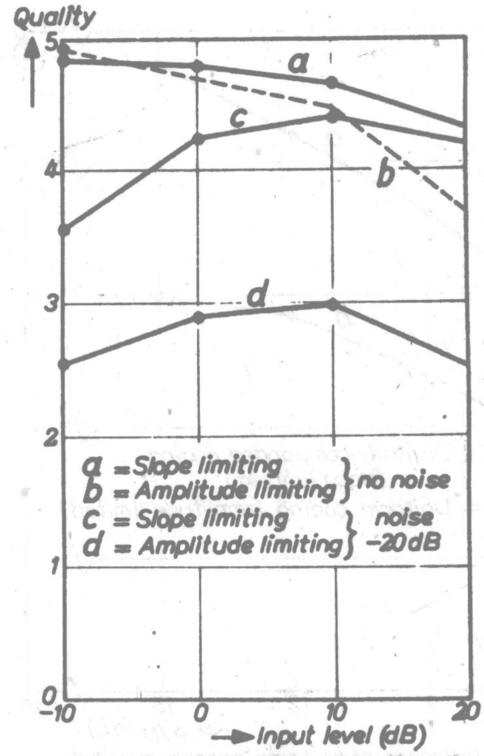

FIG.3 Speech quality as function of input level

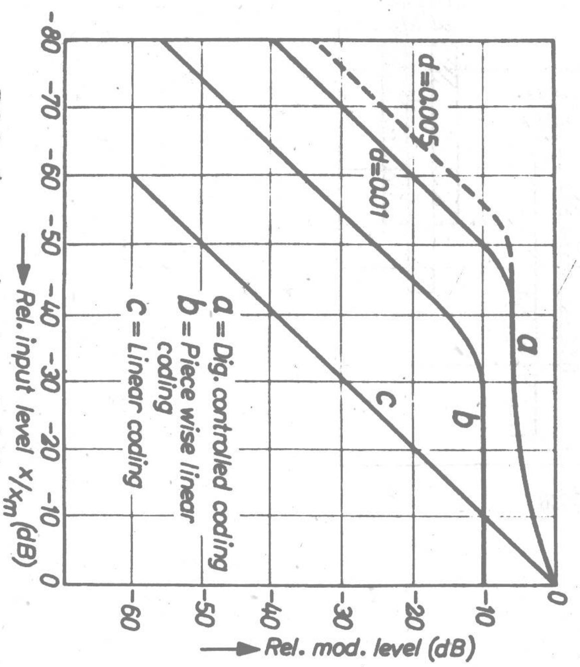

FIG.2 Compression characteristics

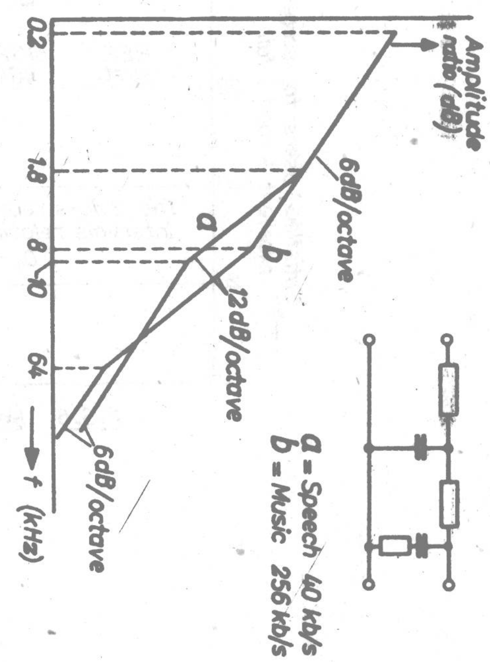

FIG.4 Characteristics of the integrating networks