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Signal-to-Noise Ratio of Pilot Tracking Tones Embedded in Binary Coded Signals

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Abstract—**Pilot tracking tones are used to derive position reference information which is employed by a head-positioning servo system to position and maintain the head precisely over a selected track in magnetic tape and disk recorders. Simple expressions are derived for the signal-to-noise ratio of pilot tracking tones embedded in a binary encoded signal.**

Introduction

THE STORAGE of digital or analog information on magnetic disk or tape is well known. Particularly common in data processing applications is the magnetic disk file in which information is written on and read from concentric tracks on the disks. At typical state-of-the-art track densities of, say, 10 tracks/mm, such a disk file must be provided with position reference information which is employed by a head-positioning servo system to position and maintain the head precisely over a selected track. The operation of maintaining the head on the track is known as "track following." In some disk files, position reference information is provided remotely from a dedicated servo disk [1]. However, at higher track densities, such an arrangement has the disadvantage that it is difficult to guarantee alignment between the remote servo disk and the actual information disk. To overcome this disadvantage, various techniques to provide position information from the information disk itself have been proposed and implemented. Servo systems for automatic track following are not only found in rather sophisticated disk files. For example, dynamic track following or automatic track finding systems are currently used in consumer-type video or digital audio tape recorders to ease the mechanical accuracies of the recorder. The techniques currently in use can be divided into two main categories:

- time-multiplex technique
- frequency-multiplex technique.

In the time-multiplex technique the signals representing the user information and the signals required for the servo systems are entirely separated in their respective places on the track. A name commonly used for this technique is sectorized servo tracking. The servo position information is a sampled signal and therefore affects the maximum attainable servo bandwidth. A compromise must be sought between the amount of servo position information and the performance of the servo system. As an example, we mention the format employed in the R-DAT recorder [2].

The frequency-multiplex technique, on the other hand, partitions, as its name suggests, the user and servo position information into separate frequency bands. This technique has the obvious advantage over the time-multiplex technique that it provides continuous information rather than sampled information. Often, the servo position information is recorded as low-frequency components, usually called pilot tracking tones [3]. In a typical embodiment the servo position information consists of two (or more) single-frequency signals recorded deeply (buried) in the magnetic medium below the area used for user information [4], [5]. The principle of operation is as follows. On even tracks the pilot tone has a frequency f_1 and on odd tracks the pilot tone frequency is f_2. Servo position information is developed from the read-back signals by subtracting the amplitude of the f_2 component from the f_1 component. These two components can be separated with band filters since they differ in frequency. As the head moves off track in one direction, the amplitude of one component decreases while the amplitude of the other increases. To circumvent interaction between the buried pilot tones and the user information, user information is often encoded in such a way that the power spectral density function of the encoded stream vanishes at the pilot tone frequencies [6].

High-density digital recording on thin metal-evaporated media employs saturation recording, i.e., two stable states of magnetization represent data to be stored. This fact precludes the obvious technique of, e.g., simply adding a sinusoidal waveform to the binary data or deeply "burying" the servo information in the medium. Pilot tracking tones, however, can be generated with the following technique [7] useful in devices such as video or audio recorders that record a stream on erased unformatted media, with no servo control of the writing process. The servo information is recorded with the user data and can then be used during reading.

As usual, user information is stored in the form of two-level (positive and negative) symbols. During a certain interval, here called positive cycle, a surplus of positive symbols is stored on the tape or disk, and in a similar

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fashion, during a negative cycle with a length equal to the positive cycle, a deficit of positive symbols is recorded. This procedure is repeated ad infinitum. Coding techniques are required to provide this degree of freedom. The pilot tone frequency can be varied by an appropriate choice of the positive and negative cycle length. The principle of operation to obtain the servo information is similar to the one previously described for the buried pilot tones. The pilot tracking tone is an intrinsic part of the binary information stream which, since the user information is a random sequence, arises from a stochastic process so that even in the instance that no noise sources are present, we conclude that the signal-to-noise ratio of the pilot tracking tone is finite. As the signal-to-noise ratio of the pilot tracking tone has its consequence on the maximum attainable bandwidth of the servo system, a compromise has to be found between the redundancy of the code and the performance of the servo system. In the subsequent sections we will assess the signal-to-noise ratio of pilot tracking tones embedded in digitally encoded signals of various coding formats.

I. Fixed Disparity Codes

It is assumed that the binary user information with a bit rate $1/T_b$, is translated into a coded sequence having a channel bit rate $1/T$, $T \le T_b$. The quotient $R = T/T_b$ is, as usual, called the rate of the code. By virtue of the physics of the recording channel, the recorded sequence consists of binary digits. Thus let a codeword $\boldsymbol{x} = (x_1, \cdots, x_n)$ consist of binary symbols, x_i, $1 \le i \le n$, $x_i \in \{-1, 1\}$. The disparity d of a codeword is the difference between the numbers of symbols with a positive or negative polarity, or

$$d = \sum_{i=1}^{n} x_i.$$

The table of codewords consists of two sets, denoted by S_+ and S_-. During the positive cycle, source words are translated into codewords selected from S_+, and in a similar way during a negative cycle the codewords are selected from S_-. The length of the positive (or negative) cycle is denoted by N. The set S_+ consists of codewords of fixed positive disparity d, $d > 0$, and the set S_- consists of codewords of fixed negative disparity $-d$. Obviously, the number N_d of codewords of length n and disparity d is given by the binomial coefficient

$$N_d = |S_+| = |S_-| = \binom{n}{(n-d)/2}.$$

The rate of the code, denoted by R, is simply

$$R = \frac{1}{n} \log_2 N_d. \tag{1}$$

The number of available codewords N_d is, in general, not a power of two, so that in practice the rate is slightly lower than given by (1).

The power spectral density function of the coded stream is the sum of a continuous and a discrete component (it is tacitly assumed that the channel symbols are transmitted serially):

$$H(fT) = H_c(fT) + H_d(fT)$$

where $H(fT)$ is the power spectral density function of the coded stream, $H_c(fT)$ and $H_d(fT)$ are the continuous and discrete components, respectively. The signal-to-noise ratio of the pilot tracking tone, denoted by SNR, is defined as the quotient of the power of the pilot tracking tone and the power of the continuous component at the pilot tracking frequency f_p, or

$$\text{SNR} = \frac{H_d(f_pT)}{H_c(f_pT)} T. \tag{2}$$

The subsequent analysis rests on the basic assumption that all codewords are chosen at random. In other words, the signal-to-noise ratio under average conditions is assessed, no attempt is made to compute the worst case performance [8].

An attractive property of the set S_+ which can easily be verified, is that the difference between the number of codewords $\boldsymbol{x}$ in S_+ with $x_i = 1$ and the number of codewords with $x_i = -1$ is fixed for all $i \in \{1, \cdots, n\}$, i.e.,

$$\frac{1}{|S_+|} \sum_{x \in S_+} x_i = u, \qquad i \in \{1, \cdots, n\}$$

is independent of the symbol position i. Similarly, we find for codewords in S_-

$$\frac{1}{|S_-|} \sum_{x \in S_-} x_i = -u, \qquad i \in \{1, \cdots, n\}.$$

With a routine enumeration we find

$$u = \frac{1}{|S_+|} \sum_{x \in S_+} x_i = \frac{d}{n}. \tag{3}$$

It can now be concluded that the discrete component is a square wave of frequency $f_p = 1/(2nNT)$ and average amplitude $u = d/n$, so that

$$H_d(fT) = \frac{8}{\pi^2} u^2 \sum_{i=0}^{\infty} \frac{1}{(2i+1)^2} \delta\{fT - (2i+1)f_pT\}$$

where $\delta(\cdot)$ is Dirac's delta function.

A. Computation of the Spectrum

The discrete parameter random process formed by the sequence of codeword symbols $\{x_i\}$ has an autocorrelation function $R(k) = E\{x_i x_{i+k}\}$, and thus is assumed to be wide-sense stationary. The average power spectral density function of the cyclostationary process of the form described earlier has been derived by Bennett [9] and further worked out in [10] and [11]. Under the assumption that the process is ergodic, the power spectral density function is given by

$$H(fT)/T = R(0) + 2 \sum_{k=1}^{\infty} R(k) \cos(2\pi k f T).$$

The computation of the power spectral density function of the code formats described in this paper is facilitated by making two important observations about the binary sequence [12].

- Besides the pilot tracking tone there is no correlation between the codewords. Therefore, if the square-wave shaped discrete component is subtracted from the coded stream, we find a signal with only a continuous power spectral density function. A further consequence of the observation is $R_c(k) = 0$, $|k| \geq n$, where $R_c(k)$ denotes the autocorrelation function of the continuous component.
- The correlation $E\{x_{j_1} x_{j_2}\}$ between two symbols in the same codeword, x_{j_1} and x_{j_2}, $j_1, j_2 \in \{1, \cdots, n\}$, $j_1 \neq j_2$, does not depend on the particular positions j_1 or j_2 of the symbols in a codeword.

This observation leads to the following reduction of the expression of the autocorrelation function of the continuous component:

$$\begin{aligned} R_c(k) &= E\{(x_j - u)(x_{j+k} - u)\} \\ &= \frac{1}{n} \frac{1}{|S_+|} \sum_{i=1}^{n-k} \sum_{x \in S_+} (x_i - u)(x_{i+k} - u) \\ &= \left(1 - \frac{k}{n}\right) \beta, \qquad 1 \leq k < n \end{aligned}$$

where

$$\beta = \frac{1}{|S_+|} \sum_{x \in S_+} (x_{j_1} - u)(x_{j_2} - u), \qquad j_1, j_2 \in \{1, \cdots, n\}, j_1 \neq j_2. \tag{4}$$

The continuous component of the power spectral density function $H_c(fT)/T$ of a sequence composed of codewords with the aforementioned properties, is given by

$$\begin{aligned} H_c(fT)/T &= R_c(0) + 2 \sum_{k=1}^{n-1} R_c(k) \cos(2\pi k f T) \\ &= R_c(0) + 2\beta \sum_{k=1}^{n-1} \left(1 - \frac{k}{n}\right) \cos(2\pi k f T) \end{aligned}$$

which can be simplified to

$$H_c(fT)/T = \alpha - \beta + \beta n F_n^2(fT) \tag{5}$$

where

$$F_n(fT) = \frac{\sin(\pi n f T)}{n \sin(\pi f T)}$$

and

$$\alpha = R_c(0) = \frac{1}{|S_+|} \sum_{x \in S_+} (x_{j_1} - u)(x_{j_1} - u).$$

Recalling that $x_i \in \{-1, 1\}$ and using (3), we find

$$\begin{aligned} \alpha &= \frac{1}{|S_+|} \sum_{x \in S_+} (x_{j_1} - u)(x_{j_1} - u) \\ &= 1 - u^2. \end{aligned} \tag{6}$$

Computation of β for Fixed Disparity Codewords

Following the definition (4) and again keeping in mind that $x_i \in \{-1, 1\}$,

$$\begin{aligned} \beta &= \frac{1}{|S_+|} \sum_{x \in S_+} (x_{j_1} - u)(x_{j_2} - u) \\ &= \frac{1}{|S_+|} \sum_{x \in S_+} x_{j_1} x_{j_2} - u^2 \\ &= \Pr(x_{j_1} = x_{j_2}) - \Pr(x_{j_1} \neq x_{j_2}) - u^2 \\ &= 2 \Pr(x_{j_1} = x_{j_2}) - 1 - u^2, \qquad u = \frac{d}{n}, j_1 \neq j_2 \end{aligned}$$

where Pr $(\cdot)$ denotes the relative number of codewords satisfying $(\cdot)$. By inspection we find

$$\Pr(x_{j_1} = 1) = \frac{1}{2}\left(1 + \frac{d}{n}\right)$$

and

$$\Pr(x_{j_1} = -1) = \frac{1}{2}\left(1 - \frac{d}{n}\right).$$

In a similar fashion we find

$$\begin{aligned} \Pr(x_{j_1} = x_{j_2} = 1) &= \frac{1}{2}\left(1 + \frac{d}{n}\right) \frac{1}{2}\left(1 + \frac{d-1}{n-1}\right) \\ \Pr(x_{j_1} = x_{j_2} = -1) &= \frac{1}{2}\left(1 - \frac{d}{n}\right) \frac{1}{2}\left(1 - \frac{d+1}{n-1}\right). \end{aligned}$$

After a routine computation, we yield

$$\begin{aligned} \beta &= 2 \Pr(x_{j_1} = x_{j_2}) - 1 - u^2 \\ &= 2 \{\Pr(x_{j_1} = x_{j_2} = 1) \\ &\quad + \Pr(x_{j_1} = x_{j_2} = -1)\} - 1 - u^2 \\ &= \frac{u^2 - 1}{n - 1}. \end{aligned}$$

A substitution, using (5) and (6), yields the following simple expression for the continuous part of the power spectral density function:

$$H_c(fT)/T = \{1 - u^2\} \frac{n}{n-1} \{1 - F_n^2(fT)\}. \tag{7}$$

Note that the shape of the power spectral density function is only a function of the codeword length. The disparity has an effect on the magnitude of the continuous and discrete components, but it does not affect their shape. Apart from the constant of proportionality and some notational difference, (7) is similar to the result derived by Franklin and Pierce [13] for zero-disparity codewords. Fig. 1 gives some examples of the power spectral density function $H_c(fT)$ of signals based on zero-disparity codewords.

At the low-frequency end, the continuous component

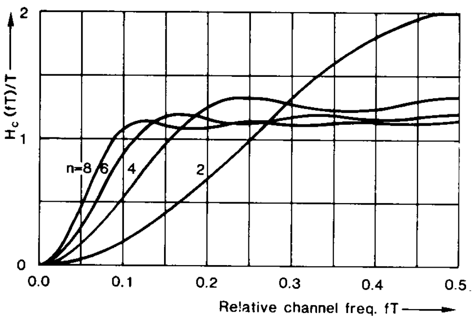

Fig. 1. Power spectral density function of zero disparity codes with codeword length n as parameter.

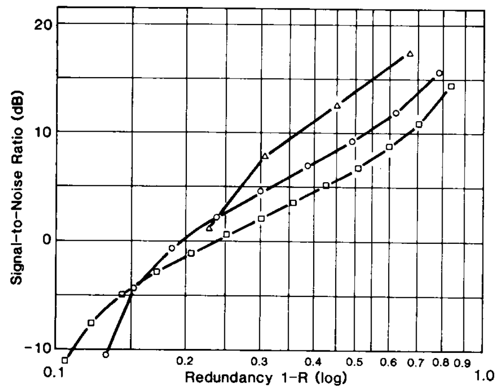

Fig. 2. SNR versus redundancy $(1 - R)$, $f_pT = 1/120$, of fixed disparity codewords. $\Delta: n = 10$, $\bigcirc: n = 20$, $\square: n = 30$.

can be approximated by the following parabola:

$$H_c(fT)/T \simeq \frac{\pi^2}{3}(1 - u^2)\, n(n + 1)(fT)^2, \quad fT \ll 1.$$

A simple substitution using (2), and further assuming that the pilot tracking tone is situated in the frequency region where the parabolic approximation of the continuous component is valid, yields

$$\mathrm{SNR} \simeq \frac{24}{\pi^4}\frac{d^2}{n^2 - d^2}\frac{1}{n(n + 1)}(f_pT)^{-2}.$$

The expression can be simplified by assuming that the codeword length n is relatively large:

$$\mathrm{SNR} \simeq \frac{24}{\pi^4}\frac{d^2}{n^4}(f_pT)^{-2}, \qquad n \gg 1.$$

We are now in the position, using (1), (2), and (7), to compute the relationship between the redundancy $(1 - R)$ of the code and the signal-to-noise ratio of the pilot tone. Fig. 2 shows the signal-to-noise ratio SNR versus the redundancy $(1 - R)$ at an arbitrarily chosen pilot tone frequency $f_pT = 1/120$, for codeword length $n = 10, 20, 30$ and disparity $2 \le d < n$. It can be observed from the figure that for a given redundancy, codes based on codewords with a small length perform slightly better than their more complex counterparts. It can further be noticed that at a redundancy $(1 - R) \simeq 0.2$, which appears to be an attractive tradeoff figure for practical systems, hardly any difference can be found.

II. Polarity Switch

We now turn to a construction of a code which is attractive as no look-up tables are required for encoding and decoding. This can be of importance in situations where speed or dissipation of the circuitry are critical parameters. A code, called polarity switch, is investigated, where $(n - 1)$ source symbols are mapped without modification onto the first $(n - 1)$ symbols of the codeword with length n. Obviously, the rate of the polarity switch code is

$$R = 1 - \frac{1}{n}. \tag{8}$$

To reduce the clerical work, we assume that n is odd. The additional nth symbol of the codeword, the so-called polarity bit, is employed to identify the polarity of the transmitted codeword. The transmitter operates on the following procedure. The polarity symbol x_n is preset to 1. The transmitter determines the disparity Σx_i of the codeword. When the transmitter is in a positive cycle, we proceed as follows. If the disparity found is positive, then the codeword is transmitted without any modification. If, on the other hand, the disparity is negative, then the entire codeword, the polarity symbol included, is inverted before transmission. A possible inversion of the codeword can be noticed at the receiver's end by simply observing the sign of the polarity bit: each received codeword is inverted if the polarity of the recovered polarity symbol is negative, and it is not inverted if that symbol is positive. During the negative cycle the polarity of the codeword symbols is chosen in such a way that the resulting disparity of the transmitted codeword is always negative. As a result of this procedure, the disparity of all transmitted codewords is positive during the positive cycle and negative during the negative cycle. When the codeword length n is even and the disparity of the codeword is nonzero, there is no difference with the previous procedure described for odd n. However, to avoid undesired discrete components in the transmitted data sequence, the transmitted codeword is at random inverted if the disparity of the codeword is zero. A similar code construction with the aim of reducing the low-frequency content was described by Greenstein [14].

The discrete component is a square wave with amplitude u, which is given by the following expression, n odd,

$$\begin{aligned} u &= \frac{1}{|S_+|}\sum_{x \in S_+} x_i = \frac{1}{2^{n-1}}\frac{1}{n}\sum_{d=1}^{n} dN_d \\ &= \frac{1}{2^{n-1}}\frac{1}{n}\sum_{d=1}^{n} d\binom{n}{(n - d)/2} \\ &= \frac{1}{2^{n-1}}\binom{n - 1}{(n - 1)/2}. \end{aligned} \tag{9}$$

Proceeding along the same lines as before, we find for n even:

$$u = \frac{1}{2^n} \binom{n}{n/2}.$$

Therefore, we conclude that the amplitude of the pilot tone decreases with increasing codeword length n. The discrete part of the power spectral density function is

$$H_d(fT) = \frac{8}{\pi^2} u^2 \sum_{i=0}^{\infty} \frac{1}{(2i+1)^2} \cdot \delta\{fT - (2i+1) f_p T\}.$$

Using the results of Section I-A, it is rather straightforward to compute the power spectral density function of the polarity switch technique. The encoder maps without modification $(n-1)$ source symbols onto the first $(n-1)$ symbols of the codeword. Assuming that the source symbols are randomly chosen, i.e., there is no correlation between the symbols, we find, using (4) and (9),

$$\beta = \frac{1}{|S_+|} \sum_{x \in S_+} (x_{j_1} - u)(x_{j_2} - u) = -u^2,$$

$$j_1, j_2 \in \{1, \cdots, n-1\}, j_1 \neq j_2.$$

To show that there is no correlation between the polarity symbol x_n and the other symbols in the codeword, that is,

$$\sum_{x \in S_+} x_n x_{j_1} = 0, \qquad j_1 \in \{1, \cdots, n-1\}$$

is somewhat more involved. The encoder presets the polarity symbol x_n to 1, thus

$$\sum_{x \in S_+} x_n x_{j_1} = \sum_{x \in S_+} x_{j_1} = 0, \qquad j_1 \in \{1, \cdots, n-1\}.$$

It can easily be seen that a possible inversion of the codeword by the encoder does not change this result. We now arrive at the following expression for the the power spectral density function of the polarity switch technique:

$$H_c(fT)/T = 1 - u^2 n F_n^2(fT). \qquad (10)$$

Fig. 3 shows the power spectral density function $H_c(fT)$ of signals encoded according to the polarity switch technique. Unlike the power spectral density function of the fixed disparity code described previously, the power spectral density function of the polarity switch system does not vanish at zero frequency. Since the pilot tone frequency is commonly situated in the low-frequency band, this fact has a detrimental effect on the signal-to-noise ratio of the pilot tone. Expression (10) can at sufficiently low frequencies be approximated by

$$H_c(fT)/T \simeq 1 - nu^2 + \frac{\pi^2}{3}(n^2 - 1) nu^2 (fT)^2, \qquad fT \ll 1$$

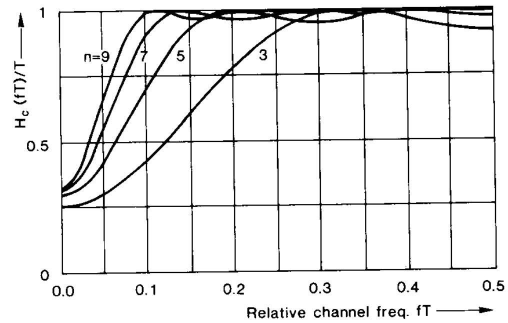

Fig. 3. Power spectral density function of polarity switch with codeword length n as parameter.

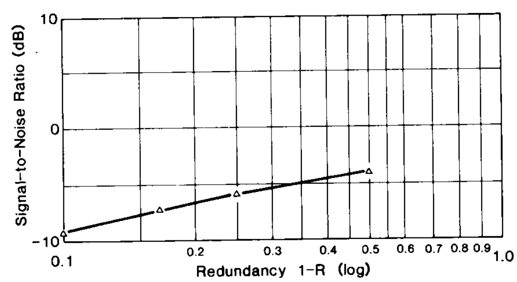

Fig. 4. SNR versus redundancy $(1 - R), f_p T = 1/120$, of polarity switch code.

which can further be simplified when the codeword length n is large:

$$H_c(fT)/T \simeq 1 - nu^2.$$

Under the mentioned conditions the signal-to-noise ratio of the pilot tracking tone is

$$\text{SNR} \simeq \frac{8}{\pi^2} \frac{u^2}{1 - nu^2}, \qquad fT \ll 1, n \gg 1.$$

Fig. 4 shows the SNR of the polarity switch technique versus the redundancy $(1 - R)$ at a pilot tone frequency $f_p T = 1/120$. Indeed, the polarity switch method has the benefit of a simple embodiment, but a comparison with Fig. 2 shows that in the entire redundancy range investigated, the signal-to-noise ratio of the pilot tone generated by the polarity switch technique is worse than that of the fixed disparity technique. For example, at a redundancy $(1 - R) \simeq 0.2$, a difference of approximately 5 dB can be observed.

Conclusion

The signal-to-noise ratio of pilot tracking tones embedded in binary coded formats has been assessed. It has been found that the signal-to-noise ratio can be improved when the redundancy of the code is increased. Since the signal-to-noise ratio of the pilot tracking tone limits the maximum attainable bandwidth of the servo system, a compromise has to be found between the redundancy of the code and the performance of the servo system. The signal-to-noise ratio of pilot tones based on the polarity switch technique, which is attractive as no look-up tables are required for encoding and decoding, is substantially smaller than that of codes based on the fixed disparity format.

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