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CCD MULTIFREQUENCY-TONE RECEIVER WITH COEFFICIENT COMPENSATION OF CHARGE-TRANSFER LOSS

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The implementation of a multifrequency-tone receiver based on the envelope detection of the received signal is presented. The envelope detection technique using a set of band-pass FIR quadrature filters, implemented by means of CCD tapped delay lines, is described. The problem of CCD charge-transfer loss is considered and a particular technique of coefficient modification is presented to compensate the above effect in the resulting frequency response. Some preliminary experimental results are finally reported by using a constructed system prototype.

1. INTRODUCTION

Multifrequency-tone transmission is widely used for automatic dialing, telemetering and data transmission in conventional telephone networks and in mobile communication systems. The signal often consists of sequences of tones chosen from a set of specified frequencies. The receiver has to determine when the signal is present and, in this case, which frequency was actually transmitted.

Many approaches have been proposed, including band-pass filtering [1] , use of Discrete Fourier Transform (DFT) [2] [3] , tone parameter extraction [4] and recently pattern recognition detection [5] . Among all the proposed techniques, one of the most promising and attractive, from a practical implementation view-point is represented by the envelope detection of the transmitted signal. Envelope detection can be performed through the multiplication of the received waveform by two in-quadrature sinusoidal waves and subsequent low-pass filtering [6] or by using two band-pass filters with a 90° degree phase difference (quadrature filters) [7] [8].

In this paper the implementation of a multifrequency-tone receiver based on the envelope detection of the received signal is presented. The envelope detection is performed by using a set of band-pass FIR quadrature filters, implemented by means of CCD tapped delay lines. After a brief recall to the complex envelope detection technique applied to the multifrequency-tone receiver, the practical CCD FIR filter implementation is described in details. In the filter design and implementation the problem of CCD charge-transfer loss is considered and a particular simple technique of coefficient modification [9] is applied to compensate for the charge-transfer loss effects on the filter frequency response. Some preliminary experimental results obtained by means of a constructed prototype are also reported, giving some indications for future improvements.

The theoretical performance of the described receiver in terms of probability of false alarm and correct detection of tones as functions of signal-to-noise ratio and receiver parameters is described in [10] , where the extension to the simultaneous transmission of more than one tone is also considered.

2. ENVELOPE DETECTION FOR A MULTIFREQUENCY-TONE RECEIVER THROUGH QUADRATURE FILTERS

The complex envelope of a real signal x(t) is defined as [11]

$$c(t) = x(t) + j\hat{x}(t) \tag{1}$$

denoting ^ the Hilbert transform operation and $|c(t)|$ the envelope of x(t).

If x(t) is a pure tone of the form

$$x(t) = A\cos(\omega t + \varphi) \tag{2}$$

it follows easily from (1) that

$$|c(t)|^2 = x^2(t) + \hat{x}^2(t) = A^2 \tag{3}$$

Let us suppose now that we receive a signal s(t) which may contain one of M possible tones having equal amplitude at the frequencies ω_i, i = 1,.. .M. In other words suppose to receive a signal of the form

$$s(t) = A\cos(\omega_i t + \varphi) + n(t) \tag{4}$$

if the tone is present, or of the form

$$s(t) = n(t) \tag{5}$$

if no tone is transmitted, being n(t) the contribution of any type of noise, interference, disturbance, that may be received superimposed on the transmitted signal.

Let us consider now M couples of band-pass quadrature filters $H_i(\omega)$ and $\hat{H}_i(\omega)$, i=1,...M. The filters $H_i(\omega)$ are non overlapping band-pass filters having ω_i as respective center frequencies. The filters $\hat{H}_i(\omega)$ are defined as

$$\hat{H}_i(\omega) = -j\mathrm{sgn}\, H_i(\omega) \qquad i=1,\dots M \qquad (6)$$

Therefore $H_i(\omega)$ and $H_i(\omega)$ represent two band-pass quadrature filters and the output of $\hat{H}_i(\omega)$ is the Hilbert transform of the output of $H_i(\omega)$. Denoting these two outputs $x_i(t)$ and $\hat{x}_i(t)$ respectively and neglecting for the moment the contribution of the term n(t) in (4) and (5) we have that either

$$x_i^2(t) + \hat{x}_i^2(t) = A^2 \qquad (7)$$

if the i-th tone ω_i was transmitted, or

$$x_i^2(t) + \hat{x}_i^2(t) = 0$$

if a different tone or no tone was transmitted.

Hence the receiver algorithm can be stated as follows:

a) determine the M envelope squares $|c_i|^2$

$$|c_i|^2 \triangleq x_i^2(t) + \hat{x}_i^2(t) \qquad i=1,\dots M \qquad (8)$$

b) choose the maximum of the M envelope squares

$$C \triangleq \max_i \{|c_i|^2\} \qquad (9)$$

which will coincide with one of the M $|c_i|^2$ as

$$C = |c_j|^2$$

c) if C>T, the j-th tone is detected, if C<T, no transmitted tone is assumed,

where T is a specified threshold that depends on the allowable probability of false alarm P_F (i.e. the probability of detecting a tone when no tone was actually transmitted). In [10] it is shown that, in presence of white Gaussian noise, the threshold T may be chosen to be

$$\frac{T}{N} = 2\log\left(\frac{M}{P_F}\right) \qquad (10)$$

where N is the noise power in the filter pass-bands (all supposed of equal bandwidth). In [10] the probability of correct detection is also shown in terms of the receiver threshold T and the signal-to-noise ratio of the received signal.

A final consideration is to be done for the implementation of the filter $H_i(\omega)$ and $\hat{H}_i(\omega)$. If we choose to implement them in discrete form, it is more convenient to consider the FIR structure instead of the IIR one. In fact the first one requires to produce the output samples of $x_i(t)$ and $\hat{x}_i(t)$, i=1,...M only at the time instants when they are necessary for the computation of the envelope squares. On the contrary the IIR structure would require the computation of all the output samples for every filter. Therefore the FIR solution allows the multiplexing of a definite hardware structure to the different filters to be implemented with considerable saving in system complexity and cost.

3. FILTER IMPLEMENTATION

The set of filters employed for frequency discrimination was implemented by means of CCD tapped delay line, due to the fact that such device is, as it is well known, well suitable for transversal filter implementation in a discrete time environment [12] .

The particular structure employed in the tone-receiver, as resulting from the above considerations, is a programmable filter in which the coefficient weights drive multiplying digital-to-analog converters (MDAC): this configuration makes possible to control the filter transfer function by means of a digital circuitry.

As it is shwon in Fig. 1, a microprocessor has the task to control the filter coefficients. The final structure results little more complicated than a classical programmable filter [9] [13] due to the fact that two sets of coefficients for the couple of band-pass quadrature filters of a given tone must be switched in a very short time (i.e. during the time interval between two sampling istants). In fact the output values of the two filters must be obtained starting from the same configuration of signal

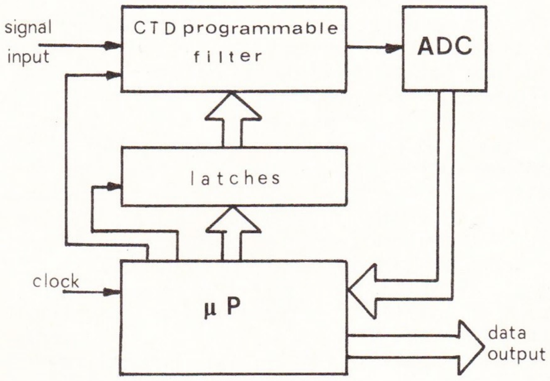

Figure 1 : System block-diagram.

samples within the CTD delay line, as it is clear from eq. (8).

Therefore it necessary to employ the external set of latches so that the microprocessor can provide the coefficients of both filters and, subsequently, only a handshake signal is required for the switching of the two sets of coefficients.

The coefficient weights of the sets of filters can further be computed in advance and stored in a suitable table so that the microprocessor can give the coefficients simply by transferring the values from the internal table to the output circuitry. Coefficient weights have been computed by means of the well known window method [14] and modified by a suitable algorithm [15] in order to compensate for the charge transfer loss which is typical of the charge-transfer-devices.

A very simple coefficient modification algorithm was indeed used. The modified coefficients h'_k are obtained from the nominal coefficients h_k through the recursive relation

$$h'_k = \frac{1}{1-k\varepsilon} \left[h_k - \varepsilon(k-1)h'_{k-1} \right] \qquad (11)$$

where ε is the value of the charge-transfer inefficiency. As shown in [9] [15], this simple algorithm achieves an improvement of the charge-transfer-loss effects on the filter frequency response by more than one order of magnitude.

4. RESULTS AND CONCLUSIONS

The receiver structure described above was realized in a prototype form, using a single charge transfer tapped delay line time-multiplexed among all the filters implemented. The number of used delays was 32. The sampling frequency was 8 KHz and M=4 tones were considered in this first protype implementation.

Two examples of the filter masks are shown in Fig. 2 and Fig. 3. They refer respectively to a band-pass filter and a band-pass Hilbert transformer having a center frequency f_c=2 KHz and a bandwidth (-6 dB) W=500 Hz. Out from a region very near to zero frequency in the Hilbert case (having only secondary effects in the present application), the response of the two filters appears to be adequate for a multifrequency-tone receiver. Better performances, of course, would be achieved by using longer tapped delay lines at the expenses however of an increase in the time required to set the filter coefficients by the microprocessor [16].

In conclusion the proposed structure appears to be well suitable for implementing a multifrequency-tone receiver. In particular it is interesting to note that the presence of a microprocessor unit allows a great system flexibility that can be exploited to perform, if necessary, some auxiliary operations. Finally a structure of the type described and implemented could be very suitable and interesting for other applications in which adaptivity is required as for the construction of adaptive equalizers.

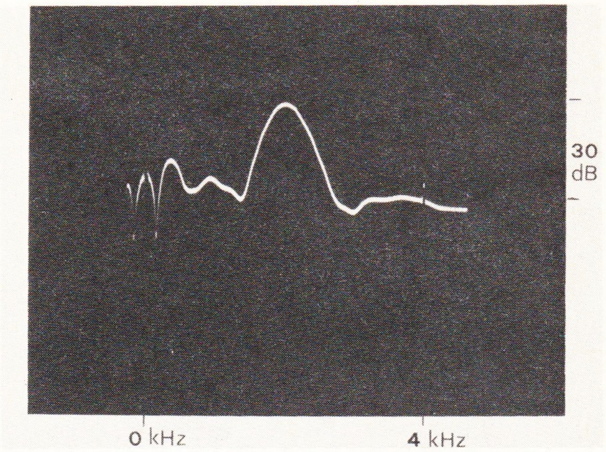

Figure 2 : Example of an implemented band-pass filter.

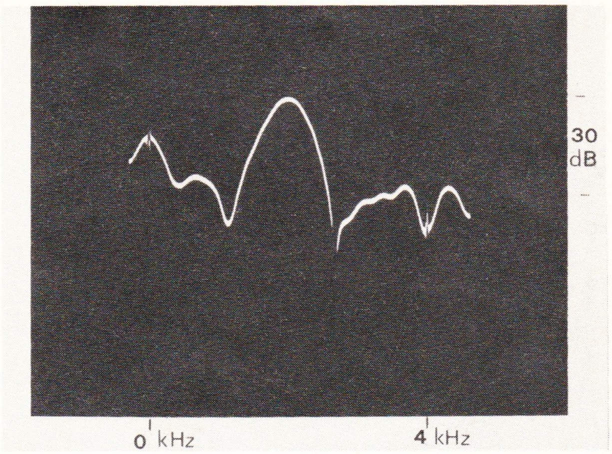

Figure 3 : Example of an implemented band-pass Hilbert transformer.

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