

Description

A. Background of the invention

A(1) Field of the invention

- 5 The invention relates to an interpolating time-discrete filter arrangement for converting a time-discrete input signal with which an input sampling frequency f_i is associated into a time-discrete output signal with which an output sampling frequency f_o is associated which is higher than the input sampling frequency.

10 A(2) Description of the prior art

As is generally known, a time-discrete signal is formed by a series of signal samples. The sampling frequency associated with such a signal indicates the rate at which these signal samples occur. The signal sample itself indicates the magnitude of the signal at a given instant. Within a certain range such a signal sample can assume any value, or only one of a number of discrete values. In the latter case a digital signal

- 15 is involved and the signal sample is usually represented by a code word having a plurality of bits. Hereinafter the signal samples of the input signal will be called input samples and will be denoted $x(q)$; $q = \dots -2, -1, 0, 1, 2, 3, \dots$. Similarly, the signal samples of the output signal will be called output samples and will be denoted $y(n)$; $n = \dots -2, -1, 0, 1, 2, 3, \dots$

Interpolating filter arrangements of the above-mentioned type have already been known for many 20 years. For the sake of brevity, for a general survey, reference is made to References 1—6 listed in paragraph C. Such arrangements produce a time-discrete output signal with which an output sampling frequency is associated with such a value that the ratio between the output sampling frequency and the input sampling frequency is a rational number. Usually, the output sampling frequency is an integral multiple of the input sampling frequency.

- 25 Practical implementations of interpolating filters are extensively described in, for example, the References 3, 4 and 5. Like all types of time-discrete filter arrangements they comprise a signal processing circuit to which the time-discrete input signal and also filter coefficients are applied. As is known, these filter coefficients represent samples of the finite impulse response of the filter and are produced by a filter coefficients generator.

30 However, in practice it has been found that there are situations in which the output sampling frequency is not a rational multiple of the input sampling frequency; it holds, for example, that $f_o = f_i \sqrt{2}$. Such a situation is found in, for example, digital audio equipment which must be intercoupled; for example a digital tuner, a digital tape recorder, a digital record-player, etc. In practice, these apparatuses each comprise their own clock generator for generating the sampling pulses required. The frequencies of these 35 clock pulse generators will never be exactly equal to each other. So as to enable the apparatuses to co-operate with each other the output sampling frequency associated with the digital signal produced by a first apparatus must be equal to the input sampling frequency accepted by the second apparatus.

An interpolation filter having a non-rational relationship between its input and output sampling rates is described in the European Patent Application EP—A—0 052 047. This prior art interpolation filter includes a 40 first circuit arrangement for producing first clock pulses at said input sampling rate, and a second arrangement for producing second clock pulses at the said output sampling rate, a coefficients generator for generating the actually required group of filter coefficients. This generator being controlled by the first and second clock pulses and comprises:—

a third circuit arrangement to which the first and second clock pulses are applied and which in 45 response to each first clock pulses produces a deviation component $d(q)$ the magnitude of which is proportional to the time interval ($T_{d(q)}$) between that first clock pulse and the immediately preceding or immediately following second clock pulse;

a fourth circuit arrangement for generating in response to the deviation component the actually required group of filter coefficients.

- 50 In this prior art interpolation filter the fourth circuit arrangement is formed as a memory in which all of the coefficients which may ever be used have been stored. These coefficients are derived from a predetermined impulse response by sampling it at an extremely high internal rate. The deviation component is used to address the memory which in response thereto supplies the actually required filter coefficients.

55 For determining the deviation component the time interval between two successive first clock pulses is partitioned into a high number of sub-intervals. The length of each sub-interval is equal to a period of the extremely high internal sampling frequency which is used to sample the impulse response. In the third circuit arrangement is determined in which sub-interval a second clock pulse is generated. The number of this sub-interval represents the said deviation component.

- 60 The accuracy of the output samples of this interpolation filter now clearly depends upon the number of sub-intervals which is fully determined by the high internal sampling frequency. So the higher the required accuracy, the higher this sampling rate. For quite normal accuracy requirements this sampling rate should be already extremely high, which makes the implementation of such interpolation filter extremely difficult 65 and expensive.

B. Object and summary of the invention

It is an object of the present invention to provide an interpolation filter arrangement for increasing the sampling rate of an input signal by a non-rational factor and which does not require the extremely high internal sampling frequency and whose output samples can have a higher accuracy.

According to the invention in this interpolation filter arrangement the third circuit arrangement for producing the deviation component comprises:

- (f) a generator (400) producing signal samples at the second sampling rate f_u which characterize a periodic signal having a period $1/f_i$ and which varies linearly between two limit values;
 - (g) selection means (404) for selecting those two consecutive signal samples produced by the generator (400) of which one is located above and the other below a predetermined threshold value;
 - (h) subtracting means (407, 408) for forming two difference components the first one of which is equal to the difference between the magnitude of the first limit value and the magnitude of the firstly selected signal sample and the second difference component being equal to the difference between the magnitude of the second limit value and the magnitude of the secondly selected signal sample;
 - (i) adding means (409) for summing the two difference components to produce a sum component;
 - (j) means (410) for dividing one of said difference components by the sum component;
- and the fourth circuit arrangement being constituted by a calculator arrangement which in response to the deviation component $d(q)$ applied thereto produces the group of W filter coefficients according to the expression

$$a(d(q), w) = h(d(q)T_u + wT_u)$$

wherein

$$w = 0, 1, 2, \dots, W-1,$$

$a(d(q), w)$ is the w^{th} filter coefficient of the group of W filter coefficients,
 $h(v)$ the impulse response of a FIR-filter,
 v a continuous variable within the interval $-\infty < v < \infty$.

The accuracy of the filter coefficients $a(d(q), w)$ now only depends upon the accuracy with which the expression can be calculated, and on the number of bits available for $d(q)$.

C. References

1. A digital Processing Approach to Interpolation; R. W. Schafer, L. R. Rabiner; Proceedings of the IEEE, Vol. 61, No. 6, June 1973, pages 692—702.
2. Arrangement for converting discrete signals into a discrete single-sideband signal frequency-multiplexed signal and vice versa; United States Patent No. 41.31.764 (PHN 8731). Paragraph E(1.2).
3. Digital filter; United States Patent No. 39.28.755 (PHN 6883).
4. Interpolating digital filter; United States Patent No. 39.88.607 (PHN 7733).
5. Interpolating digital filter with input buffer; United States Patent No. 39.97.773 (PHN 7729).
6. Interpolation-Decimation Circuit for Increasing or Decreasing Digital Sampling Frequency; R. E. Crochiere, L. R. Rabiner; United States Patent 4,020,332.
7. Theory and Application of Digital Signal Processing; L. R. Rabiner; B. Gold.
8. Terminology in Digital Signal Processing; L. R. Rabiner, et al; IEEE Transactions on Audio and Electroacoustics, Vol. AU-20, December 1972, pages 322—337.

D. Short description of the figures

Figure 1 shows the theoretically functional structure of an interpolating time-discrete filter arrangement;

Figure 2 shows the finite impulse response of the "analog" filter used in the filter arrangement shown in Figure 1;

Figure 3 shows some time diagrams to explain the filter arrangement of Figure 1;

Figure 4 shows an embodiment of the interpolating time-discrete filter arrangement according to the invention;

Figure 5 shows a time diagram to explain the operation of the filter coefficients generator used in the embodiment shown in Figure 4;

Figure 6 shows a different form of the impulse response.

Figure 7 shows a modification which can be made in the filter coefficients generator shown in Figure 4, for the case in which the impulse response has the form shown in Figure 6;

Figure 8 shows a time-discrete sawtooth generator for use in the filter coefficients generator shown in Figure 4;

Figure 9 shows some time diagrams to explain a further embodiment of the filter coefficients generator;

Figure 10 and Figure 11 show embodiments of the signal processing arrangement for use in the filter arrangement shown in Figure 4.

E. Description of the embodiments

E(1) Theoretical background

Figure 1 shows schematically the theoretical model of a time-discrete filter arrangement for changing the sampling frequency of a time-discrete input signal. It comprises an "analog" filter 1 followed by a sampling arrangement 2. The time-discrete input signal formed by the series of input samples $x(q)$, $q = \dots, -2, -1, 0, 1, 2, 3, \dots$ which occur at the input sampling frequency f_i is applied to this filter 1. This filter supplies an output signal $z(t)$ which is equal to the sum of all the so-called individual output signals $z_{x(q)}(t)$. The response of this filter 1 to the input sample $x(q)$ produces such an individual output signal. As is known, this individual output signal is thus equal to the product of this input sample and the impulse response of this filter. If this impulse response is represented by the function $h(v)$, this individual output signal can be expressed mathematically as follows:

$$z_{x(q)}(t) = x(q)h(t - t_{x(q)}) \quad (1)$$

in which $t_{x(q)}$ represents the instant at which the input sample $x(q)$ occurs. Let it be assumed that for this instant it holds that:

$$\begin{aligned} t_{x(q)} &= qT_i \\ T_i &= 1/f_i \end{aligned} \quad (2)$$

so that

$$z_{x(q)}(t) = x(q)h(t - qT_i) \quad (3)$$

The impulse response $h(v)$ is defined for all values of v , but it will be assumed that it assumes values unequal to zero only at finite intervals. Therefore this filter 1 is sometimes called an FIR-filter (=Finite Impulse Response-filter). Let it be assumed that this impulse response has the shape shown in Figure 2. If now the input samples $x(q)$ occur at the instants qT_i as they are shown at A in Figure 3, then all the individual output signals are known. Some of these individual output signals, namely those for $q = -3, -2, -1, 0, 1, 2$ are shown at B, C, D, E, F, G in Figure 3. Herein it is assumed that $x(q) = 1$ for all values of the independent variable q . As mentioned in the foregoing, the actual output signal $z(t)$ of this filter 1 is formed by the mathematical sum of all of the individual output signals. Since the impulse response is finite, the number of individual output signals contributing to the ultimate output signal $z(t)$ is also finite. Let it be assumed that of only M individual output signals this contribution is unequal to zero, that the filter is a casual filter and that the input signal sample applied last to the filter has the number q , then it holds that:

$$z(t) = \sum_{m=0}^{M-1} z_{x(q-m)}(t) \quad (4)$$

$$t < (q+1)T_i$$

This output signal and thus any of the individual output signals is now sampled in the sampling arrangement 2 by, for example, the sequence of sampling pulses indicated at H in Figure 3 and occurring at the output sampling frequency f_u , which is here chosen to be equal to $2f_i\sqrt{2}$. The n^{th} sampling pulse will be denoted $ku(n)$ and occurs at an instant $t_{ku(n)}$ which is defined by the expression:

$$t_{ku(n)} = t_0 + nT_u \quad (5)$$

Herein t_0 represents an arbitrary time constant and it will also be assumed that $0 \leq t_0 < T_u$. Thus, the sampling arrangement 2 produces a series of output samples, the output sample obtained in response to the n^{th} sampling pulse $ku(n)$ being denoted by $y(n)$ and being defined by the expression:

$$y(n) = z(t_0 + nT_u) = \sum_{m=0}^M x(q-m)h(t_0 + nT_u - (q-m)T_i) \quad (6)$$

If more specifically f_u is smaller than f_i , then a decimating filter arrangement is involved. If in contrast therewith f_u exceeds f_i an interpolating filter arrangement is involved. The following description will be based on interpolating filter arrangements and it will be assumed that $f_u = Rf_i$, wherein $R \geq 1$. In this connection this quality R is sometimes denoted as an interpolation factor.

Because with an interpolating filter arrangement the output sampling frequency f_u is higher than the input sampling frequency, a plurality of output samples occur in the time interval T_i located between the two consecutive input samples $x(q)$ and $x(q+1)$. The number of the first output sample occurring in that time interval will be denoted by n_q and the r^{th} output sample in that interval by n_{q+r} . If, further, the distance between the first output sample $y(n_q)$ in this interval and the immediately preceding input sample $x(q)$ is denoted by $T_{d(q)}$, then it follows from expression (6) that:

$$y(n_q+r) = \sum_{m=0}^{M-1} x(q-m)h(T_{d(q-m)} + (n_q - n_{q-m} + r)T_u) \quad (7)$$

5 Herein:

$$T_{d(q-m)} = t_0 + n_{q-m}T_u - (q-m)T_1 \quad (8)$$

10 In expression (7) the quantities $h(T_{d(q-m)} + wT_u)$ wherein $w = (n_q - n_{q-m} + r)$, represent the filter coefficients of the time-discrete filter arrangement. In the following paragraph these filter coefficients will be denoted for the sake of brevity by $a(d(q), w)$. The quantity $d(q)$ will be called the deviation component and is defined as follows:

$$d(q) = T_{d(q)}/T_u$$

15 E(2) Some special values for R

In the preceding paragraph it was assumed that in principle the interpolation factor may have any random positive value exceeding or equal to unity. In this paragraph some special cases will be further described.

20 In the first case it is assumed that $R=1$ so that $f_u = f_i$ and consequently $T_u = T_1$, it then further holds that:

$$r=0$$

$$q-m = n_{q-m}$$

$$T_{d(q-m)} = t_0 \quad (10)$$

$$n_q - n_{q-m} = mT_1$$

$$n_q = q$$

So that each output sample is determined by the expression:

$$y(q) = \sum_{m=0}^{M-1} x(q-m)h(t_0 + mT_1) \quad (11)$$

Embodiments of time-discrete filter arrangements whose operation is fully defined by expression (11) are extensively described in chapter 9 of Reference 7 and in Reference 8, as well as in many other publications.

40 In the second case it is assumed that $R > 1$. Now a distinction can be made between the case in which R is an integer and the case in which R is not an integer.

If R is an integer then it holds that:

$$r=0, 1, 2, \dots, R-1$$

$$T_1 = RT_u \quad (12)$$

$$T_{d(q)} = t_0$$

50 So that each output sample is now defined by the expression:

$$y(Rq+r) = \sum_{m=0}^{M-1} x(q-m)h(t_0 + rT_u + mT_1) \quad (13)$$

55 Embodiments of such interpolating time-discrete filter arrangements are extensively described in, for example, References 1, 3 and 4.

For the case that an interpolation factor R is to be realised which is not an integer but is rational, so that it can be written as a quotient of two integers, for example $R=L/P$, a decimating time-discrete filter arrangement having a decimation factor P can be arranged in cascade with an interpolating time-discrete filter arrangement having an interpolation factor L. Interpolating time-discrete filter arrangements of this type are described in, for example, References 3, 5 and 6.

60 The situation is quite different if the interpolation factor R is not an integer and is also irrational. In that case the number of output samples occurring between two consecutive input samples is not always the same. Let it be assumed that this number is not more than R' , then a group of a total of $W=MR'$ filter

coefficients $a(d(q), w)$ must be available for calculating all the output samples. In accordance with expression (7), these filter coefficients are then obtained from the expression:

$$a(d(q), w) = h(T_{d(q)} + wT_u) \quad (14)$$

$$w = 0, 1, 2, \dots, MR'$$

As $T_{d(q)}$ now also changes from input sample to input sample, the filter coefficients of the group also change their values continuously. Embodiments of interpolating time-discrete filter arrangements with an irrational interpolation factor will be described in the following paragraph.

E(3) Interpolating time-discrete filter arrangement with an irrational interpolation factor

Figure 4 shows schematically an embodiment of an interpolating time-discrete filter arrangement having an irrational interpolation factor. In known manner it comprises a signal processing arrangement 3 and a filter coefficients generator 4. This signal processing arrangement 3 is preceded by a buffer store 5 to which the input samples $x(q)$ are applied. Such an input sample $x(q)$ is stored in this buffer store at the instant at which an input clock pulse $ki(q)$ occurs at its write input WR. At the instant at which an output clock pulse $ku(n)$ is applied to the read input R of this buffer store, the content of this buffer store is applied to the signal processing arrangement 3. Let it now be assumed that this buffer store is reset as a result thereof. If now the subsequent output clock pulse $ku(n)$ occurs before a new input clock pulse $ki(q)$ has occurred, then this buffer store supplies a signal sample having the value zero. The input clock pulses $ki(q)$ are produced by a clock pulse generator 6 and occur with the input sampling frequency f_i . The output clock pulses $ku(n)$ are produced by a clock pulse generator 7 and occur with the output sampling frequency f_u .

As is obvious from expression (7), the filter coefficients $a(d(q), w)$, cannot be calculated until the time interval $T_{d(q)}$ and consequently the deviation component $d(q)$ are known. As has already been mentioned, this deviation component represents the ratio between the time interval $T_{d(q)}$ located between the input sample $x(q)$ and the immediately subsequent output sample $y(n_q)$ and the output sampling period T_u or, which is the same, the ratio between the time interval located between the instant $t_{ki(q)}$ at which an input clock pulse $ki(q)$ occurs and the instant $t_{ku(n_q)}$ at which an immediately subsequent output clock pulse $ku(n_q)$ occurs and the interval T_u between two consecutive output pulses.

In the filter coefficients generator 4 shown in Figure 4 this deviation component $d(q)$ is calculated in a particularly efficient way and with reference to the associated group of W filter coefficients. The generator comprises a, preferably digital, sawtooth generator 400 which produces at a rate f_u digitally encoded samples of a periodic analog sawtooth-shaped signal which has a period T_i . For a better understanding, as shown in Figure 4, the generator 400 may be assumed to comprise a sawtooth generator 400(1) which is controlled by the input clock pulses $ki(q)$. This generator produces, for example, the analog sawtooth-shaped signal shown at A in Figure 5, which signal varies between the values $+E$ and $-E$ and its value suddenly changes from $+E$ into $-E$ at the instant an input clock pulse $ki(q)$ occurs. For the sake of completeness, these input clock pulses are shown at B in Figure 5. The analog sawtooth signal thus obtained is thereafter sampled in a sampling device 400(2) by the output clock pulses $ku(n)$ at the instants $t_{ku(n)}$, as they are shown, for example, at C in Figure 5. This sampling device 400(2) produces the signal samples $s(n)$ which are shown at A in Figure 5 by means of arrows and are encoded digitally in an analog-to-digital converter 400(3). It will now be explained in greater detail how the deviation component $d(q)$ can be determined.

Let it be assumed that a given input clock pulse $ki(q)$ occurs between the two consecutive output clock pulses $ku(n_{q-1})$ and $ku(n_q)$, the distance between these two last-mentioned clock pulses being T_u . The distance between $ki(q)$ and $ku(n_q)$ is the time interval $T_{d(q)}$ looked for. If now the signal samples which can be taken from the sawtooth signal at the instants $t_{ku(n_{q-1})}$ and $t_{ku(n_q)}$ have the respective values $s(n_{q-1})$ and $s(n_q)$, then it follows that

$$T_{d(q)} : (T_u - T_{d(q)}) = (E - |s(n_q)|) : (E - |s(n_{q-1})|)$$

so that

$$d(q) = \frac{E - |s(n_q)|}{(E - |s(n_q)|) + (E - |s(n_{q-1})|)} = \frac{E - |s(n_q)|}{O(n_q)} \quad (15)$$

From this it follows that the deviation component $d(q)$ can be fully determined from the value of the signal samples of the sawtooth signal. As is further shown in Figure 4, these signal samples $s(n)$ are applied for this purpose to a cascade arrangement of two shift register elements 401 and 402 and to a shift register element 403. The contents of these shift register elements are shifted under the control of the output clock pulses $ku(n)$. The signal samples $s(n)$ are also applied to a zero-crossing detector circuit 404 which produces a detection pulse each time a signal sample of the positive polarity is followed by a signal sample of the negative polarity. This detection pulse is applied to the clock pulse inputs of two further shift register elements 405 and 406. The signal inputs thereof are connected to the signal outputs of the respective shift register elements 402 and 403. In response to this detection pulse the contents of the shift register elements 405 and 406 become equal to $s(n_{q-1})$ and $s(n_q)$, respectively. In subtracting arrangements 407 and 408 the

absolute value of these signal samples is subtracted from the value E. The two difference components thus obtained are added together in an adder arrangement 409 and the sum component $O(n_q)$ thus obtained is divided in a divider stage 410 by the difference component produced by subtracting arrangement 407. The deviation component $d(q)$ thus obtained is applied to an arithmetic unit 411 (for example a microcomputer) which is arranged to calculate at the given value of $d(q)$, the required filter coefficients $a(d(q),w)$ (four in the case described) in accordance with expression (14).

E(4). Special embodiments

As in the coefficients generator whose embodiment is shown in Figure 4, no restriction whatsoever is imposed on the shape of the impulse response $h(v)$, it is advantageous to use a microcomputer 411 to calculate the $W(=4)$ filter coefficients, starting from the calculated value of the deviation component $d(q)$. Special purpose hardware may advantageously be used when specially shaped impulse responses are defined, for example when the impulse response has the shape shown in Figure 6 and which is defined as follows:

$$\begin{aligned} h(v) &= 0 & \text{for } v \leq 0 \text{ and } v \geq 2T_u \\ h(v) &= \frac{v}{T_u} H & \text{for } 0 < v \leq T_u \\ h(v) &= (2 - \frac{v}{T_u}) H & \text{for } T_u < v < 2T_u \end{aligned} \quad (16)$$

Herein H represents a constant. From this it then follows that:

$$\begin{aligned} a(d(q), 0) &= d(q)H \\ a(d(q), 1) &= (1 - d(q))H \\ a(d(q), 2) &= 0 \\ a(d(q), 3) &= 0 \end{aligned} \quad (17)$$

Then, $d(q)$ follows from expression (15) and it further holds that:

$$1 - d(q) = \frac{E - s(n_q - 1)}{(E - s(n_q)) + (E - |s(n_q - 1)|))} = \frac{E - s(n_q - 1)}{O(n_q)}$$

Put differently, in this case the factor $1 - d(q)$ can be obtained by dividing the output signal of the subtracting arrangement 408 (see Figure 4) by the sum component $O(n_q)$. For the sake of completeness, all this is shown schematically in Figure 7. As is shown in this Figure, the output signal of the subtracting arrangement 408 is applied to a divider stage 412, which also receives the sum component $O(n_q)$ from adder arrangement 409. The output components of the two divider stages 410 and 412 are multiplied by the constant H in multiplier stages 413 and 414.

As in this case the sum component $O(n_q)$ is present in both $d(q)$ and $1 - d(q)$, one may alternately proceed as follows. Choose the filter coefficients as follows:

$$\begin{aligned} a(d(q), 0) &= E - s(n_q) \\ a(d(q), 1) &= E - s(n_q - 1) \\ a(d(q), 2) &= 0 \\ a(d(q), 3) &= 0 \end{aligned}$$

The signal processing arrangement 3 now supplies output samples $y'(n)$ which are applied to a multiplying circuit in which they are multiplied by $H/O(n_q)$, as a result of which the desired output samples $y(n)$ can be obtained.

In the coefficients generator shown in Figure 4 the signal samples $s(n)$ are obtained by sampling an analog sawtooth signal. A sawtooth generator which produces these signal samples in a fully digital way is shown in Figure 8. The generator comprises an input circuit which includes a D flip-flop 414. The D-input thereof continuously receives the logic value "1". The clock input CL receives the input clock pulses $ki(q)$ and the reset input R receives a reset signal. The Q-output of this D flip-flop 414 is connected to the D-input of a D flip-flop 415. The output clock pulses $ku(n)$ are applied to the clock pulse input CL of this flip-flop. The pulses occurring at the Q-output of the D flip-flop 415 and the output clock pulses $ku(n)$ inverted in an inverter 416 are applied to a NAND-gate 417 the output pulses of which are applied as reset pulses to the reset input R of D flip-flop 414. This input circuit formed by two D flip-flops may, for example, be the dual D flip-flop of the "74" series marketed by Signetics. The output pulses of D flip-flop 415 are applied to a

phase-locked loop (PLL) which is of a digital construction. This loop comprises an up/down counter 418 whose counting range is located in the range 0—2E. The output pulses of D-flip-flop 415 are applied to its up-counting input (+). The counting position of this counter 418 is applied with intervals T_u to a digital low-pass filter 420 via a subtracting arrangement 419. In the subtracting arrangement 419 the counting position is reduced by a reference number REF which, for example, equal to half (E) of the range (2E) of counter 418.

The low-pass filter 420 which has a bandwidth of less than 1 Hz, produces at a rate f_u output numbers which are applied to an accumulator formed by an adder 421 and a delay network 422 whose time delay is equal to T_u . Adder 421 has two outputs denoted by c and e, respectively. At the output c the (most significant) carry bit occurs of the word produced by the adder, whilst the remaining bits of this word occur at the output e. The output c is further connected to the down-counting input (−) of the up/down counter 418 and the output e represents at the same time the output of the sawtooth generator.

In the foregoing it was assumed that generator 400(1) produces a sawtooth signal. It is, however, alternatively possible to construct this generator such that it produces a signal which varies triangularly in the way shown in, for example, Figure 9. To determine the deviation component $d(q)$ the two signal samples $s(n_q+2)$ and $s(n_q+3)$ may be the starting point. It should be noted that this is also possible for the case shown schematically in Figure 5 where a sawtooth-shaped signal is involved. In that case the zero-crossing detector 404 must produce an output pulse for each signal sample having a negative polarity followed by a signal sample having a positive polarity.

The filter coefficients generator 4 operates as follows. The time interval T_u between two consecutive output clock pulses $ku(n)$ is distributed in an infinite number of sub-intervals T_{uo} which are infinitely small. Thereafter it is determined between which two consecutive output clock pulses $ku(n)$ the input clock pulse $ki(q)$ is located and also in which sub-interval. The number of this sub-interval may be considered to be the deviation component $d(q)$. The number of sub-intervals now depends on the number of bits by which $d(q)$ is represented. If this number of bits is chosen such that a countable number of sub-intervals is obtained, the following procedure can be followed. The MR' filter coefficients $a(d(q),w)$ associated with each value of the deviation component can be calculated in advance and stored in a storage medium. Let it be assumed that the deviation component can assume sixteen different values, then this storage medium contains, for example, 16 MR' filter coefficients. A selector to which the deviation component is applied and which depending on the value thereof selects the desired group from these sixteen groups of filter coefficients. In such cases the term "table look up" is sometimes used.

E(5). Embodiments of the signal processing arrangement

A particularly suitable embodiment of the signal processing arrangement is known from Figure 9.10 of Reference 7 and is shown in Figure 10, for the sake of completeness. It comprises in known manner a number of MR' multiplying circuits $30(\cdot)$ (4 in the embodiment shown), a plurality of adder circuits $31(\cdot)$ and a plurality of delay circuits $32(\cdot)$, each having a time delay T_u . All the multiplying circuits are connected by means of an input to the input 33 of the signal processing arrangement and thus receive the input samples $x(q)$ simultaneously. In addition they receive via a filter coefficients inputs the required filter coefficients produced by the filter coefficients generator shown in Figure 4. The desired output samples now occur at the output 35 of adder circuit $31(2)$.

An alternative embodiment of the signal processing arrangement 3 is shown for the greater part in Figure 9.1 of Reference 7 and, for the sake of completeness, is shown in Figure 11. It comprises a cascade arrangement of a number of delay devices $36(\cdot)$ (three in the embodiment shown), each having a time delay T_u . This cascade arrangement is connected to the input 33 of the signal processing arrangement. Inputs and outputs of the delay devices $36(\cdot)$ are connected via multiplying circuits $30(\cdot)$ to inputs of an adder device 37, which supplies the desired output samples $y(n)$ from its output 35. The multiplying circuits $30(\cdot)$ receive the required filter coefficients via their filter coefficients inputs. In the case in which all the filter coefficients associated with a given value of the deviation component are supplied simultaneously by the filter coefficients generator 4 as suggested in Figure 5, the filter coefficients $a(d(q),0)$ are directly applied to the multiplying circuit $30(0)$, and the filter coefficients $a(d(q),1)$ are applied to the multiplying circuit $30(1)$ via a delay device $38(1)$ having a time delay T_u . Similarly, the filter coefficients $a(d(q),2)$ and $a(d(q),3)$ are applied to the respective multiplying circuits $30(2)$ and $30(3)$ via delay devices $38(2)$ and $38(3)$ having the respective time delays $2T_u$ and $3T_u$.

E6. General remarks

In the foregoing it was assumed that the buffer store 5 (see Figure 4) produces an input sample having the value zero for each output clock pulse produced, unless this output clock pulse occurs immediately after an original input sample had occurred. In practice, it has however been found that it is alternatively possible to assign the value 1 to the said input sample instead of the value zero; however, this requires a modification of the original impulse response.

In the foregoing, calculating the deviation component $d(q)$ was based on the time interval located between an input clock pulse $ki(q)$ and the immediately preceding output clock pulse $ku(n)$. It is, however, also permissible to define this deviation component as the quotient between the time interval located

between an input clock pulse $ki(q)$ and the immediately preceding output clock pulse $ku(n-1)$ and the time interval T_u between two consecutive output clock pulses.

Claims

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1. An interpolation discrete time filter arrangement for increasing the sampling rate of a discrete time input signal from f_i to f_u , wherein f_u is a non-rational multiple (R) of f_i , this discrete-time input signal being constituted by a series of input components $x(q)$, wherein $q=..., -2, -1, 0, 1, 2, ...,$ comprising:

- (a) a filter input for receiving the input signal;
- 10 (b) first means for producing first clock pulses $ki(q)$ at the input sampling rate f_i ;
- (c) second means for producing second clock pulses $ku(q)$ at the output sampling rate f_u ;
- (d) a filter coefficients generator for producing a group of W filter coefficients, this generator being controlled by the first and the second clock pulses and comprising:
 - d1) means to which the first clock pulses ki and the second clock pulses ku are applied and which in
 - 15 response to each q^{th} first clock pulse produces a deviation component $d(q)$ the magnitude of which is proportional to the time interval $T_{d(q)}$ between the q^{th} first clock pulse $ki(q)$ and the immediately preceding or the immediately following second clock pulse;
 - d2) means for generating in response to the deviation component $d(q)$ the group of W filter coefficients;
- 20 (e) a processor arrangement coupled to the filter input and to the filter coefficients generator for multiplying input components $x(q)$ with subsets of filter coefficients of the group of W filter coefficients and accumulating the products obtained by multiplying the input components by a particular subset of filter coefficients; characterized in that the means for producing the deviation component comprises:
 - (f) a generator (400) producing signal samples at the second sampling rate f_u which characterize a
 - 25 periodic signal having a period $1/f_i$ and which varies linearly between two limit values;
 - (g) selection means (404) for selecting those two consecutive signal samples produced by the generator (400) of which one is located above and the other below a predetermined threshold value;
 - (h) subtracting means (407, 408) for forming two difference components the first one of which is equal to the difference between the magnitude of the first limit value and the magnitude of the firstly selected
 - 30 signal sample and the second difference component being equal to the difference between the magnitude of the second limit value and the magnitude of the secondly selected signal sample;
 - (i) adding means (409) for summing the two difference components to produce a sum component;
 - (j) means (410) for dividing one of said difference components by the sum component;
 - in that the generating means (411) being constituted by a calculator arrangement which in response to
 - 35 the deviation component $d(q)$ applied thereto produces the group of W filter coefficients according to the expression

$$a(d(q),w)=h(q)T_u+wT_u)$$

wherein

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$$w=0, 1, 2, ..., W-1,$$

$a(d(q),w)$ is the w^{th} filter coefficient of the group of W filter coefficients,
 $h(v)$ the impulse response of a FIR-filter,
 v a continuous variable within the interval $-\infty < v < \infty$.

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2. Arrangement as claimed in Claim 1, characterized in that the said periodic signal has the shape of a sawtooth.

3. Arrangement as claimed in Claim 1 and 2, characterized in that the two limit values are positive and negative respectively and that the said threshold value is the zero value.

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Patentansprüche

- 1. Eine interpolierende zeitdiskrete Filteranordnung zum Vergrössern der Abtastgeschwindigkeit eines zeitdiskreten Eingangssignals von f_i zu f_u , wobei f_u ein nicht rationelles Vielfach (R) von f_i ist, wobei das
- 55 zeitdiskrete Eingangssignal durch eine Reihe von Eingangsabtastwerten $x(q)$ gebildet wird, wobei $q=..., -2, -1, 0, 1, 2, ...$ ist, mit:
 - (a) einem Filtereingang zum Erhalten des Eingangssignals;
 - (b) ersten Mitteln zum Erzeugen erster Taktimpulse $ki(q)$ mit der Eingangsabtastgeschwindigkeit f_i ;
 - (c) zweiten Mitteln zum Erzeugen zweiter Taktimpulse $ku(q)$ mit der Ausgangsabtastgeschwindigkeit f_u ;
 - 60 (d) einem Filterkoeffizientengenerator zum Erzeugen einer Gruppe von W Filterkoeffizienten, wobei dieser Generator durch die ersten und die zweiten Taktimpulse gesteuert wird und mit:
 - d1) Mitteln, denen die ersten Taktimpulse ki und die zweiten Taktimpulse ku zu zugeführt werden und die in Antwort auf jeden q . ersten Taktimpuls einen Abweichungsanteil $d(q)$ erzeugen, dessen Grösse dem
 - 65 Zeitintervall $T_{d(q)}$ zwischen dem q . ersten Taktimpuls $ki(q)$ und dem unmittelbar vorhergehenden oder dem unmittelbar nachfolgenden zweiten Taktimpuls proportional ist;



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between an input clock pulse $k_i(q)$ and the immediately preceding output clock pulse $k_u(n-1)$ and the time interval T_u between two consecutive output clock pulses.

Claims

- 5 1. An interpolation discrete time filter arrangement for increasing the sampling rate of a discrete time input signal from f_i to f_u wherein f_u is a non-rational multiple (R) of f_i , this discrete-time input signal being constituted by a series of input components $x(q)$, wherein $q = \dots, -2, -1, 0, 1, 2, \dots$, comprising:
- 10 (a) a filter input for receiving the input signal;
(b) first means for producing first clock pulses $k_i(q)$ at the input sampling rate f_i ;
(c) second means for producing second clock pulses $k_u(q)$ at the output sampling rate f_u ;
(d) a filter coefficients generator for producing a group of W filter coefficients, this generator being controlled by the first and the second clock pulses and comprising:
- 15 d1) means to which the first clock pulses k_i and the second clock pulses k_u are applied and which in response to each q^{th} first clock pulse produces a deviation component $d(q)$ the magnitude of which is proportional to the time interval $T_{d(q)}$ between the q^{th} first clock pulse $k_i(q)$ and the immediately preceding or the immediately following second clock pulse;
d2) means for generating in response to the deviation component $d(q)$ the group of W filter coefficients;
- 20 (e) a processor arrangement coupled to the filter input and to the filter coefficients generator for multiplying input components $x(q)$ with subsets of filter coefficients of the group of W filter coefficients and accumulating the products obtained by multiplying the input components by a particular subset of filter coefficients; characterized in that the means for producing the deviation component comprises:
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25 periodic signal having a period $1/f_i$ and which varies linearly between two limit values;
(g) selection means (404) for selecting those two consecutive signal samples produced by the generator (400) of which one is located above and the other below a predetermined threshold value;
(h) subtracting means (407, 408) for forming two difference components the first one of which is equal to the difference between the magnitude of the first limit value and the magnitude of the firstly selected
30 signal sample and the second difference component being equal to the difference between the magnitude of the second limit value and the magnitude of the secondly selected signal sample;
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(j) means (410) for dividing one of said difference components by the sum component;
in that the generating means (411) being constituted by a calculator arrangement which in response to
35 the deviation component $d(q)$ applied thereto produces the group of W filter coefficients according to the expression

$$a(d(q), w) = h(q)T_u + wT_u$$

wherein

$$w = 0, 1, 2, \dots, W-1,$$

$a(d(q), w)$ is the w^{th} filter coefficient of the group of W filter coefficients,

$h(v)$ the impulse response of a FIR-filter,

v a continuous variable within the interval $-\infty < v < \infty$.

- 45 2. Arrangement as claimed in Claim 1, characterized in that the said periodic signal has the shape of a sawtooth.

3. Arrangement as claimed in Claim 1 and 2, characterized in that the two limit values are positive and negative respectively and that the said threshold value is the zero value.

Patentansprüche

- 50 1. Eine interpolierende zeitdiskrete Filteranordnung zum Vergrössern der Abtastgeschwindigkeit eines zeitdiskreten Eingangssignals von f_i zu f_u , wobei f_u ein nicht rationelles Vielfach (R) von f_i ist, wobei das zeitdiskrete Eingangssignal durch eine Reihe von Eingangsabtastwerten $x(q)$ gebildet wird, wobei
55 $q = \dots, -2, -1, 0, 1, 2, \dots$ ist, mit:
- (a) einem Filtereingang zum Erhalten des Eingangssignals;
(b) ersten Mitteln zum Erzeugen erster Taktimpulse $k_i(q)$ mit der Eingangsabtastgeschwindigkeit f_i ;
(c) zweiten Mitteln zum Erzeugen zweiter Taktimpulse $k_u(q)$ mit der Ausgangsabtastgeschwindigkeit f_u ;
60 (d) einem Filterkoeffizientengenerator zum Erzeugen einer Gruppe von W Filterkoeffizienten, wobei dieser Generator durch die ersten und die zweiten Taktimpulse gesteuert wird und mit:
- d1) Mitteln, denen die ersten Taktimpulse k_i und die zweiten Taktimpulse zu zugeführt werden und die in Antwort auf jeden q . ersten Taktimpuls einen Abweichungsanteil $d(q)$ erzeugen, dessen Grösse dem Zeitintervall $T_{d(q)}$ zwischen dem q . ersten Taktimpuls $k_i(q)$ und dem unmittelbar vorhergehenden oder dem
65 unmittelbar nachfolgenden zweiten Taktimpuls proportional ist;

d2) Mitteln zum in Antwort auf den Abweichungsanteil $d(q)$ Erzeugen der Gruppe von W Filterkoeffizienten;

(e) einer mit dem Filtereingang und dem Filterkoeffizientengenerator gekoppelten Bearbeitungsanordnung zum Multiplizieren von Eingangsabtastwerten $x(q)$ mit Teilgruppen von 5 Filterkoeffizienten der Gruppe von W Filterkoeffizienten und zum Addieren der durch Multiplikation der Eingangsabtastwerte mit einer bestimmten Teilgruppe von Filterkoeffizienten erhaltenen Produkte; dadurch gekennzeichnet, dass

die Mittel zum Erzeugen des Abweichungsanteils die folgenden Elemente aufweisen;

(f) einen Generator (400), der Signalabtastwerte mit der zweiten Abtastgeschwindigkeit f_u erzeugt, die 10 ein periodisches Signal mit einer Periode $1/f_i$ kennzeichnen und die zwischen zwei Grenzwerten linear variiert;

(g) Selektionsmittel (404) zum Selektieren der zwei aufeinanderfolgenden von dem Generator (400) erzeugten Signalabtastwerte, von denen der eine über und der andere unter einem vorbestimmten Schwellenwert liegt;

15 (h) Subtrahiermittel (407, 408) zum Bilden zweier Differenzanteile, von denen der erste der Differenz zwischen der Grösse des ersten Grenzwertes und der Grösse des zuerst selektierten Signalabtastwertes entspricht und wobei der zweite Differenzanteil der Differenz zwischen der Grösse des zweiten Grenzwertes und der Grösse des an zweiter Stelle selektierten Signalabtastwertes entspricht;

(i) Addiermittel (409) zum Summieren der zwei Differenzanteile zum Erzeugen eines Summenanteils;

20 (j) Mittel (410) zum Teilen eines der Differenzanteile durch den Summenanteil;

die Erzeugungsmittel (411) durch eine Zählordnung gebildet sind, die in Antwort auf den ihr zugeführten Teilungsanteil $d(q)$ die Gruppe von W Filterkoeffizienten erzeugt und zwar entsprechend dem Ausdruck:

$$25 \quad a(d(q), w) = h(q)T_u + wT_w$$

wobei

$$w = 0, 1, 2, \dots, W-1,$$

30 $a(d(q), w)$ der w . Filterkoeffizient der Gruppe von W Filterkoeffizienten,

$h(v)$ die Impulsstossantwort eines FIR-Filters und v eine kontinuierliche Veränderliche in dem Intervall $-\infty < v < \infty$ ist.

2. Anordnung nach Anspruch 1, dadurch gekennzeichnet, dass das periodische Signal sägezahnförmig ist.

35 3. Anordnung nach Anspruch 1 und 2, dadurch gekennzeichnet, dass die zwei Grenzwerte positiv bzw. negativ sind und der Schwellenwert Null ist.

Revendications

40 1. Dispositif de filtre d'interpolation discret dans le temps pour augmenter la fréquence d'échantillonnage d'un signal d'entrée discret dans le temps de f_i à f_u , f_u étant un multiple non rationnel (R) de f_i , ce signal d'entrée discret dans le temps étant constitué d'une série de composantes d'entrée $x(q)$, où $q = \dots, -2, -1, 0, 1, 2, \dots$, comprenant:

(a) une entrée de filtre pour recevoir le signal d'entrée;

45 (b) un premier dispositif pour produire des premières impulsions d'horloge $ki(q)$ à la fréquence d'échantillonnage d'entrée f_i ;

(c) un second dispositif pour produire des secondes impulsions d'horloge $ku(q)$ à la fréquence d'échantillonnage de sortie f_u ;

(d) un générateur de coefficients de filtrage pour produire un groupe de W coefficients de filtrage, ce 50 générateur étant commandé par les premières et les secondes impulsions d'horloge et comprenant:

d1) un dispositif auquel les premières impulsions d'horloge ki et les secondes impulsions d'horloge ku sont appliquées et qui, en réaction à chaque $q^{\text{ième}}$ première impulsion d'horloge, produit une composante de déviation $d(q)$ dont la grandeur est proportionnelle à l'intervalle de temps $T_{d(q)}$ entre la $q^{\text{ième}}$ première 55 impulsion d'horloge $ki(q)$ et la seconde impulsion d'horloge immédiatement précédente ou immédiatement suivante;

d2) un dispositif pour produire, en réponse à la composante de déviation $d(q)$, le groupe de W coefficients de filtrage;

(e) un dispositif de traitement couplé à l'entrée de filtre et au générateur de coefficients de filtrage pour multiplier des composantes d'entrée $x(q)$ par des sous-jeux de coefficients de filtrage du groupe de W 60 coefficients de filtrage et pour accumuler les produits obtenus par multiplication des composantes d'entrée par un sous-jeu particulier de coefficients de filtrage; caractérisé

en ce que le dispositif pour produire la composante de déviation comprend:

(f) un générateur (400) produisant des échantillons de signal à la seconde fréquence d'échantillonnage f_u qui caractérisent un signal périodique ayant une période $1/f_i$ et qui varie linéairement entre deux valeurs 65 limites;



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(g) un dispositif de sélection (404) pour sélectionner les deux échantillons de signal successifs produits par le générateur (400) dont l'un est situé au-dessus et l'autre, en dessous d'une valeur de seuil prédéterminée;

5 (h) un dispositif soustracteur (407, 408) pour former deux composantes de différence, dont la première est égale à la différence entre la grandeur de la première valeur limite et la grandeur de l'échantillon de signal sélectionné en premier lieu et dont la seconde est égale à la différence entre la grandeur de la seconde valeur limite et la grandeur de l'échantillon de signal sélectionné en second lieu;

(i) un dispositif additionneur (409) pour sommer les deux composantes de différence en vue de produire une composante de somme;

10 (j) un dispositif (410) pour diviser une des composantes de différence par la composante de somme; en ce que le dispositif générateur (411) est constitué par un calculateur qui, en réaction à la composante de déviation $d(q)$ qui y est appliquée, produit le groupe de W coefficients de filtrage selon l'expression:

15
$$a(d(q), w) = h(d(q)T_u + wT_u)$$

où

$$w = 0, 1, 2, \dots, W-1,$$

20 $a(d(q), w)$ est le $w^{\text{ième}}$ coefficient de filtrage du groupe de W coefficients de filtrage.

$h(v)$ est la réponse impulsionnelle d'un filtre FIR (à réponse impulsionnelle de durée finie), v est une variable continue dans l'intervalle $-\infty < v < \infty$.

2. Dispositif suivant la revendication 1, caractérisé en ce que le signal périodique a la forme d'une dent de scie.

25 3. Dispositif suivant les revendications 1 et 2, caractérisé en ce que les deux valeurs limites sont négative et positive respectivement et en ce que la valeur de seuil est la valeur zéro.

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