Device and method for processing a real subband signal for reducing aliasing effects

Description

The present invention relates to processing audio or video signals and, in particular, to filter banks for converting a signal to a spectral representation, the spectral representation comprising bandpass signals or spectral coefficients.

In applications in consumer and communication electronics, the possibility of manipulating the signal spectra by frequency-selectively increasing or decreasing signal intensities is desirable, such as, for example, for equalizer functionalities or echo suppression. In connection with audio coding methods which are based on spectrally decomposing the input signal, it seems to be obvious to correspondingly amplify and/or attenuate the decoded spectral components (subband samples and/or transform coefficients) by means of multiplication by amplification factors which may be temporally alterable. The filter banks used here, however, usually have real values and involve critical sampling. Consequently, the subband signals contain aliasing components which, however, compensate one another with a decomposition (analysis) and merging (synthesis) performed directly one after the other, but not after a manipulation described above. The result may be audible interferences, such as, for example, by amplitude modulation.

Fig. 5 shows a filter bank system comprising an analysis filter bank 50 and a synthesis filter bank 51. A discrete-time signal x(n) is fed to a number of N bandpass filters 53 to obtain bandpass signals decimated by one decimator 54 per filter bank channel. The decimated bandpass signals $x_0(m)$ to $x_{N-1}(m)$ are then fed to an equalizer stage 55, wherein a special weighting coefficient g_0, g_1, ..., g_{N-1} is

associated to each bandpass signal. The weighted bandpass signals y_0 to y_{N-1} are then each fed to an interpolator 56 and filtered by a respective synthesis filter g_0, g_1, g_{N-1} 57. The filtered signals are then added by means of an adder 58 to obtain the output signal y(n) at an output 59 of the filter bank. The signal y(n) is identical to the signal x(n) when all the amplification factors g_0, g_1, ..., g_{N-1} = 1 and when the filters 53 and 57 are adjusted such that the filter bank is of a perfectly reconstructing characteristic.

It is to be pointed out that the filters h_0 are typically cosine-modulated versions of a prototype low-pass filter and that the synthesis filters g_0, g_1, ..., g_{N-1} are also correspondingly modulated versions of a prototype filter, wherein a filter g_i is matched to a filter h_i so that no artefacts are introduced by filtering.

There are filter banks comprising, for example, 20 filter bank channels, such that 20 samples of the signal x(n) result in one sample of a subband signal x_i each. In this case, the filter bank is referred to as being maximally decimated. Typically, filter banks are realized by numerically efficient mathematical methods such that the filtering occurring in each channel and subsequent decimating are performed in one processing step so that there is no undecimated signal anywhere. Alternative implementations, however, are also known and are implemented depending on the requirements.

When such a filter bank has very many channels, such as, for example, 1024 channels, this represents a transform. The transform rule so to speak implements filtering and decimating "at one strike". An MDCT having 1024 samples may thus be described by the analysis filter bank part 50 in Fig. 5, wherein N in this case would be 1024 and a single sample of a "subband signal" would be generated per block of samples fed to such a transform. The time form of a

subband signal would then result if there were several MDCT blocks of spectral coefficients and if the value of the MDCT coefficients in the subsequent blocks for a frequency index were taken as a time signal. Subsequently, whenever subbands and subband values are mentioned, both subband filtering and a transform are meant, without explicitly pointing out every time that a transform represents maximally decimated subband filtering, wherein the number of channels N equals the number of transform coefficients.

The critical sampling of the filter bank signals used has the result that bandpass filters have an overlapping region, i.e. that exemplarily the upper half of the pass region of the filter h_0 overlaps the lower half of the neighboring filter h_1. At the same time, the upper region of the filter $h_1(n)$ overlaps the lower region of the next filter $h_2(n)$. If there is a signal component in this overlapping region, both the subband signal $x_0(m)$ and the subband signal $x_1(m)$ will have information on this signal component. If both subbands are amplified identically by the amplification factors g_0 and g_1, that is both amplification factors are equal, the aliasing will be cleared up again by the analysis filters g_0, g_1 and the subsequent addition of the output signals of the filters g_0 and g_1 so that the output signal $y(n)$ is free of artefacts. However, if the two signals x_0, x_1 are amplified differently, the portion of the signal components in the overlapping region will also be amplified differently, the result of which is that the receive signal $y(n)$ has more aliasing since the synthesis filter bank "has not expected" a different weighting for the two overlapping filters.

Such artefacts can be avoided by using complex-value filter banks which, however, do not comprise critical sampling and are thus not suitable for encoding. On the other hand, such complex-value filter banks are used in post-processing units, such as, for example, in bandwidth extension (SBR) and in parametric multi-channel encoding (BCC/EBCC).

One possible but expensive solution for this problem is offered by real-value synthesis filtering including subsequent complex-value analysis, manipulation and complex-value synthesis. The implementation expenses of this approach can be reduced considerably by approximating the sequential application of real-value synthesis and complex-value analysis by so-called multiband filtering for generating the imaginary parts (''r2i'') necessary. After a corresponding re-conversion into real parts (''i2r''), the conventional real-value synthesis can be applied.

Such a complicated solution of the aliasing problem using a complex filter bank implementation is shown in Fig. 6. The real signal present in a subband representation is converted to a real time signal by means of a real synthesis filter bank, as is illustrated in Fig. 6 as output signal 61 of the real synthesis filter bank 60. This real time signal 61 is then fed to a complex analysis filter bank 62 to obtain complex-value subband signals 63. These complex subband signals 63 are then fed to a manipulation stage 64 which receives the weighting factors c_k and/or weighting factors g_i of Fig. 5 and which may be set up in the same way as the equalizer stage 55 of Fig. 5. At the output of the manipulation stage 64, there are manipulated complex-value subband signals 65 which are then converted to an output signal 67 which is again a real signal by means of a complex synthesis filter bank 66. In order to obtain a real signal 67, in the complex synthesis filter bank 66, a real part is formed either before adding, which may be identical to the adding of the adder 58 of Fig. 5, or a real part is formed after the addition 58. The imaginary part is simply discarded. The processing by means of the complex analysis filter bank/synthesis filter bank device, however, ensures that there are no more aliasing interferences in the output signal 67 caused by a different manipulation of neighboring subbands.

However, this solution is, as has been mentioned, very complicated since, in contrast to a direct manipulation of Fig. 5, an additional complex analysis filter bank and an additional complex synthesis filter bank are required, these filter banks being complicated from a calculating point of view and resulting in a delay since the analysis filters and/or synthesis filters comprise considerable filter lengths which may really be in a range of more than 30 filter taps per filter bank, i.e. may be on both the analysis side and the synthesis side. On the other hand, no aliasing is introduced.

A solution which is less complicated is the multiband filtering shown in Fig. 7, where temporally and/or frequency-neighboring subband signals are united such that the processing stages 62, 66 of Fig. 6 are no longer required but replaced by the r2i and i2r blocks, respectively, of Fig. 7. The manipulation by the manipulation stages 64 or 55 would thus take place in the complex representation, i.e. between stages 70 and 71 at 72 in Fig. 7.

In a real-to-complex (r2c) conversion, the series connection of the real-value synthesis filter bank and the complex-value analysis filter bank has been approximated. Here, an imaginary part for each real subband sample is formed by overlaying three filter output signals. The three filters are each applied in the respective subband and the two neighboring bands.

Correspondingly, the conversion from complex to real (c2r) approximates the series connection of the complex-value synthesis filter bank and the real-value analysis filter bank. Here, the real part is formed as a mean value of the original real subband sample and the overlaying of three filter output signals. The three filters are each applied to the imaginary parts in the respective subband and in the two neighboring bands.

The series connection of r2c and c2r has to reconstruct the original subband signal as precisely as possible to avoid audible interferences in the output signal. Thus, the corresponding filters must have relatively great lengths.

Such a multiband filtering which can be applied in an advantageous manner is disclosed in German patent DE 102 34 130 B3. A filter bank device for generating a complex spectral representation of a discrete-time signal includes means for generating a block-wise real-value spectral representation of the discrete-time signal, the spectral representation comprising temporally successive blocks, each block comprising a set of real spectral coefficients. Additionally, means for post-processing the block-wise real-value spectral representation is provided to obtain a block-wise complex approximated spectral representation comprising successive blocks, each block comprising a set of complex approximated spectral coefficients, wherein a complex approximated spectral coefficient may be represented by a first sub-spectral coefficient and a second sub-spectral coefficient, wherein at least either the first or the second sub-spectral coefficient can be established by combining at least two real spectral coefficients. The first sub-spectral coefficient is a real part of the complex approximated spectral coefficient and the second sub-spectral coefficient is an imaginary part of the complex approximated spectral coefficient. The combination is a linear combination and the means for post-processing is implemented to combine, for determining a complex spectral coefficient of a certain frequency, a real spectral coefficient of the frequency and a real spectral coefficient of a neighboring higher or lower frequency or a corresponding real spectral coefficient of a current block, a temporally preceding block or a temporally successive block.

This procedure is of disadvantage in that relatively long filters are necessary to obtain a representation free of aliasing, in particular in the case that there is no different weighting in two neighboring subbands, but so to speak the "benign" case that both subbands are weighted equally. If too short filters are used, aliasing will occur also in the benign case, which is unacceptable and results in great filter lengths in blocks r2i 70 and/or i2r 71 in Fig. 7. Great filter lengths, however, at the same time mean calculating complexity and, in particular, a great delay, which is also undesirable for certain applications.

It is the object of the present invention to provide an efficient and nevertheless high-quality concept for processing signals.

This object is achieved by a device for processing a real subband signal according to claim 1, a method for processing a real subband signal according to claim 24, an analysis filter bank according to claim 25, a synthesis filter bank according to claim 26, a method for analyzing a signal according to claim 27, a method for synthesizing a signal according to claim 28 or a computer program according to claim 29.

The present invention is based on the finding that the problems of high artefacts or, in exchange for this, great filter lengths can be solved by dividing subband filtering to a normal portion including a subband weighted by a weighting factor for a special subband, and a correction portion depending on another subband and a weighting factor for the other subband. Thus, an efficient implementation is achieved, since long filters for generating a normal portion are no longer necessary according to the invention, since this normal portion is no longer filtered "back and forth", but simply weighted. The aliasing problem is solved using the correction portion calculated in dependence on another subband which will, for example, be the neighboring

subband, and in dependence on a weighting factor for this subband. In order to render the inventively weighted subband "aliasing-proof", the correction portion is combined with the weighted subband, such as, for example, added, to obtain a corrected subband signal resulting in reduced aliasing.

The present invention is of advantage in that the main portion of the corrected subband signal is calculated directly, i.e. only using weighting of the weighting factor provided for this subband signal, and not subjected to filtering. This saves calculating time and, at the same time, calculating power, which is especially important, in particular for mobile or battery-driven apparatuses. No more filtering of the subband itself by means of a long analysis filter and subsequent synthesis filter is necessary. Instead, only sample after sample has to be weighted. This is achieved by a division into the weighted portion and the correction term.

The correction term may additionally be calculated by considerably shorter filters, since the precision of the correction portion does not have to be as high as the precision of the normal portion. Inventively, any scaling can be achieved in that filters for the correction portion can be made longer, thereby further reducing aliasing, but, in particular when small aliasing is still allowable, that very short filter lengths can be used for calculating the correction portions which in the extreme case may even be degenerated to a pure multiplication by a factor and subsequent addition, wherein in this case aliasing will of course be stronger than in the case of a filter length of, for example, more than 8 filter taps. On the other hand, it is ensured according to the invention by the subdivision to a normal contribution and a correction contribution that, when there is a benign subband processing, i.e. when neighboring subbands are to be weighted by the same weighting factor, no interference will be introduced

although only short filters are used. In this case, the correction portion can simply be set to zero according to the invention, which may be performed manually, which, however, is also achieved automatically when the correction term of a weighting is weighted by the difference of the weighting factors for two neighboring subbands. If the difference equals zero, the two amplification factors will be equal, i.e. the correction term will be set to zero, independent of how big it is, independent of whether the correction term would have provided only a coarse correction or even a fine correction. If the correction term had been calculated more precisely, for this case of identical weighting factors, it would have had to be zero all by itself. With only a coarse correction, however, in this case too, the result would be a correction term unequal to zero which, however, does not corrupt the result since weighting has been performed using the difference of the correction factors.

Depending on the implementation, the correction term is not made only with one subband signal, but with two neighboring subband signals and/or with as many neighboring subband signals as band overlapping within a significant region takes place with, i.e. in the region of the filter where the filter has an attenuation of, for example, smaller than 30 dB. If more than three filters overlap, more than three filters will be considered when calculating the correction term and, of course, also the correction factors associated therewith.

It is to be pointed out that the inventive concept cannot only be performed for filter banks having a relatively small number of channels, where the subband signals are bandpass signals. Rather, the inventive concept may also be applied to filter banks having a great number of filter bank channels, such as, for example, filter banks implemented by a transform. Such a transform is, for example, an FFT, a DCT, an MDCT or another maximally

decimated transform where one spectral coefficient per filter bank channel is generated per block of samples. The spectral coefficients having the same coefficient index from a sequence of temporally sequential blocks of spectral coefficients represent a bandpass signal which can be filtered for the purpose of determining a correction term to obtain a correction term.

Preferred embodiments of the present invention will be detailed subsequently referring to the appended drawings, in which:

Fig. 1 is a block circuit diagram of an inventive device for processing a real subband signal according to a preferred embodiment of the present invention;

Fig. 2 is a detailed illustration of the correction term determiner of Fig. 1;

Fig. 3a is a schematic illustration of the inventive device according to a preferred embodiment of the present invention;

Fig. 3b is a more detailed illustration of the filter part of Fig. 3a;

Fig. 3c is a schematic illustration of the inventive device according to an alternative embodiment of the present invention;

Fig. 3d is a more detailed illustration of the device schematically shown in Fig. 3c;

Fig. 4 shows an analysis filter bank/synthesis filter bank device having a subband-wise device for processing;

Fig. 5 shows a real-value analysis/synthesis filter bank device having an equalizer stage;

Fig. 6 shows cascading of a real synthesis filter bank with a complex analysis filter bank and a complex synthesis filter bank;

Fig. 7 is a schematic illustration of multiband filtering;

Fig. 8 is a more detailed illustration of filter operations for the multiband filtering of Fig. 7;

Fig. 9 is a tabular illustration of the filters for subband signals having even and odd indices;

Fig. 10 shows an exemplary comparison of magnitude frequency responses for filters for determining a correction term;

Fig. 11 shows subband filtering of an impulse;

Fig. 12 shows subband filtering of a sine tone at 1% above the band limit;

Fig. 13 shows subband filtering of a sine tone at 5% above the band limit;

Fig. 14 shows subband filtering of a sine tone at 10% above the band limit;

Fig. 15 shows subband filtering of a sine tone at 20% above the band limit;

Fig. 16 shows subband filtering of a sine tone at 30% above the band limit;

Fig. 17 shows subband filtering of a sine tone at 40% above the band limit; and

Fig. 18 is a schematic illustration of the aliasing reduction for an MDCT with a sine tone at 10% above the band limit.

Fig. 1 shows an inventive device for processing a real subband signal x(k) of a plurality of real subband signals which are an illustration of a real discrete-time signal x(n) generated by an analysis filter bank (50 in Fig. 5). The inventive device includes a weighter 10 for weighting the subband signal x_k by a weighting factor c_k determined for the subband signal to obtain a weighted subband signal 11. The weighter is preferably implemented to perform a multiplication. In particular, subband samples which are samples of a bandpass signal or spectral coefficients of a transform spectrum are multiplied by the correction factor. Alternatively, instead of multiplication, addition of logarithm values may also be performed, namely an addition of the logarithm of the correction value and the logarithm of the subband sample x_k.

The inventive device for processing further includes a correction term determiner for calculating a correction term, the correction term determiner being implemented to calculate the correction term using at least another subband signal x_l and using another weighting factor c_l which is provided for the other subband signal, the other weighting factor differing from the weighting factor c_k. This differentiation of the two weighting factors is the cause for aliasing in a real filter bank application, even when analysis and synthesis filters have a perfectly reconstructing characteristic. The correction term at the output of the means 12 is fed to a combiner 13 as is the weighted subband signal, the combiner being implemented to combine the weighted subband signal and the correction term to obtain a corrected subband signal y_k.

The combiner 13 is preferably implemented to perform a combination sample by sample. Thus, there is a "correction term sample" for every sample of the weighted subband signal x_k such that a 1:1 correction can be performed. Alternatively, however, for implementations less complicated as to calculating, a correction can be performed such that, for example, a single correction term is calculated for a certain number of weighted subband samples, which is then added in a smoothed or an unsmoothed manner to every sample of the group of samples associated to the correction term sample. Depending on the implementation, the correction term can also be calculated as a factor and not as an additive term. The combiner would in this case perform a multiplication of a correction term by the weighted subband signal to obtain a corrected subband signal y_k.

It is to be pointed out that aliasing occurs when two subband signals have been generated by filters having overlapping pass characteristics. In special filter bank implementations, there are such overlapping filter characteristics comprising an overlapping region which is significant for neighboring subband signals.

Preferably, the correction term determiner is thus implemented as is illustrated in Fig. 2. The correction term determiner includes a first correction term determiner portion 12a and a second correction term determiner portion 12b. The first correction term determiner portion considers the overlapping of the current subband signal of the index k and the next higher subband signal of the index k+1. Additionally, the correction term determiner portion 12a, apart from the subband signal x_{k+1}, also receives the weighting factor c_{k+1} of the higher subband signal. Preferably, the correction term determiner will also receive the difference of c_{k+1} and c_k, which in Fig. 2 is illustrated by q_k.

The second correction term determiner portion 12b considers the overlapping of the subband signal x_k with the subband signal x_{k-1} lower by 1 with regard to its index. The correction term determiner portion 12b thus, apart from the subband signal x_{k-1}, also receives the weighting factor c_{k-1} for this subband and preferably the difference of the weighting factor c_{k-1} and the weighting factor c_k, which in Fig. 2 is referred to by c_k.

On the output side, the first correction term determiner portion 12a provides a first correction term $q_k \times u_k$ and the second correction term determiner portion 12b provides a second correction term $p_k \times L_k$, wherein these two correction terms are added to then be combined with the weighted subband signal $c_k \times x_k$, as will be described referring to Figs. 3a and 3b.

A preferred implementation which is shown in greater detail in Figs. 8 and 3a will be detailed subsequently.

The series connection of a real-value synthesis filter bank and a complex-value analysis filter bank is approximated in multiband filtering. Here, an imaginary part for each real subband sample is formed by overlapping three filter output signals. The three respective filters are applied in the respective subband and in the two neighboring bands.

Correspondingly, the conversion from complex to real (c2r) approximates the series connection of a complex-value synthesis filter bank and a real-value analysis filter bank. Here, the real part is formed as a mean value of the original real subband sample and the overlapping of three filter output signals. The three respective filters are applied to the imaginary parts in the respective subband and the two neighboring bands.

The series connection of r2c and c2r has to reconstruct the original subband signal as precisely as possible in order to avoid audible interferences in the output signal. Thus, the corresponding filters must have relatively great lengths.

The approach presented here is based on the idea of subdividing the series connection of "r2c", "gain control" and "c2r" into signal portions forming when using equal amplification factors, and signal portions forming due to differences between amplification factors of neighboring subbands.

Since the first signal portion is to correspond to the original subband signal, the respective operation can be omitted.

The remaining signal portions are dependent on the differences of the respective amplification factors and only serve the reduction of aliasing components, as would take place in the usual r2c and c2r conversion. Since the respective filters do not influence the reconstruction of unchanged subband signals, they may comprise considerably shorter lengths.

Subsequently, the procedure will be described in greater detail.

The imaginary part in the subband k is calculated from the real-value subband samples of the subbands k, k-1 and k+1 to form:

$$I_k(z) = H_m(z)X_k(z) + H_u'(z)X_{k-1}(z) + H_l'(z)X_{k+1}(z). \quad (1)$$

The differentiations between H and H′ are necessary due to the mirroring of the subbands having odd indices.

If every subband is multiplied each by an amplification factor c_k, the result for the reconstructed signal in the subband k, considering an additional normalization factor of 0.5, will be:

$$Y_k(z) = 0{,}5\left(c_k X_k(z) + c_k G_m(z) I_k(z) + c_{k-1} G_u'(z) I_{k-1}(z) + c_{k+1} G_l'(z) I_{k+1}(z)\right)$$

If c_{k-1} is replaced by $c_k + p_k$, with $p_k = c_{k-1} - c_k$, and if c_{k+1} is replaced by $c_k + q_k$, with $q_k = c_{k+1} - c_k$, the result will be :

$$\begin{aligned} Y_k(z) = {} & 0{,}5 c_k \left(X_k(z) + G_m(z) I_k(z) + G_u'(z) I_{k-1}(z) + G_l'(z) I_{k+1}(z)\right) \\ & + 0{,}5\left(p_k G_u'(z) I_{k-1}(z) + q_k G_l'(z) I_{k+1}(z)\right). \end{aligned} \tag{2}$$

Here, the first term corresponds to the subband signal which is reconstructed when using the same amplification factors in all subbands, and thus equals the original subband signal except for the factor c_k, and/or should be like that. However, the second term represents the influence of different amplification factors and can be considered as a correction term for the subband k of the complex processing compared to the real processing. It is calculated as follows:

$$\begin{aligned} C_k(z) = {} & 0{,}5 p_k G_u'(z)\left(H_m'(z) X_{k-1}(z) + H_u(z) X_{k-2}(z) + H_l(z) X_k(z)\right) \\ & + 0{,}5 q_k G_l'\left(H_m'(z) X_{k+1}(z) + H_u(z) X_k(z) + H_l(z) X_{k+2}(z)\right). \end{aligned} \tag{3}$$

The following connections result from the characteristics of the polyphase filter bank and the mirroring of the subbands having odd indices:

$$\begin{aligned} & H_m'(z) = -H_m(z), H_l'(z) = -H_u(z), H_u'(z) = -H_l(z), \\ & G_m(z) = -H_m(z), G_u'(z) = H_l(z), G_l'(z) = H_u(z), \\ & G_u'(z) H_u(z) = H_l(z) H_u(z) \approx 0, G_l'(z) H_l(z) = H_u(z) H_l(z) \approx 0. \end{aligned} \tag{4}$$

A substitution has the following result:

$$C_k(z) = 0{,}5 p_k H_l(z)\left(H_l(z)X_k(z) - H_m(z)X_{k-1}(z)\right) + 0{,}5 q_k H_u(z)\left(H_u(z)X_k(z) - H_m(z)X_{k+1}(z)\right). \quad (5)$$

Since the reconstruction is no longer dependent on the filters used with amplification factors constant over the subbands, they can be replaced by shorter ones, wherein a respective product filter can also be approximated so that two correction terms can then be calculated instead of the imaginary part:

$$L_k(z) = 0{,}5\left(H_{ll}(z)X_k(z) - H_{lm}(z)X_{k-1}(z)\right)$$

$$U_k(z) = 0{,}5\left(H_{uu}(z)X_k(z) - H_{um}(z)X_{k+1}(z)\right)$$

$$with\ H_{ll}(z) \approx H_l^2(z), H_{lm}(z) \approx H_l(z)H_m(z), H_{uu}(z) \approx H_u^2(z), H_{um}(z) \approx H_u(z)H_m(z). \quad (6)$$

The desired subband signal including aliasing compensation is obtained by a weighted overlapping of the original subband signal and the two correction signals:

$$Y_k(z) = c_k X_k(z) + p_k L_k(z) + q_k U_k(z). \quad (7)$$

However, in practical realizations it must be kept in mind that a delay compensating the delay in the respective signal paths including filtering has to be introduced into the signal paths without filtering.

For checking the overall performance, the following pictures show the output signals after filter bank analysis, attenuation of a subband by 20 dB and subsequent filter bank synthesis for different input signals.

The approach described may also be combined with the MDCT instead of the filter bank used in EBCC.

Suitable filter coefficients for filters having a length of 5 have been generated for this. This corresponds to the uncut filters resulting when sequentially applying the corresponding transforms and/or re-transforms. Compared to the "r2c-c2r" technology, however, the new method is of advantage in that it does not generate approximation errors as long as the MDCT spectrum remains unchanged. With "r2c-c2r", however, errors would result since only the two respective neighboring bands are considered in the approximation.

The resulting signal spectra for a sine tone which is 10% above a band limit show that the aliasing components are also reduced very efficiently in connection with MDCT. Here, too, the neighboring band has been attenuated by 10 dB.

Thus, the equalizer functions and/or echo suppression methods can be integrated directly in an audio decoder, such as, for example, MPEG-AAC, by the inverse MDCT before re-transforming.

Fig. 8 shows a schematic illustration of the filter operations for real to complex (r2c) and complex to real (c2r). The imaginary component I_k of the band x_k is generated by a subband signal x_{k-1} filtered by the filter H'_u and by the subband signal of subband x_{k+1} generated by the filter H'_1. In addition, a component of the subband signal x_k filtered by the filter H_m contributes to the imaginary component I_k. Because the portion of the subband signal x_{k-1} overlapped by the filter k has a low-pass characteristic, the filter H'_u is a low-pass filter. In analogy, the portion of the upper subband signal x_{k+1} overlapped by the filter for x_k is a high-pass signal, so that H'_1 is a high-pass filter. As has already been

explained, H and H′ are differentiated to consider the mirroring of the subbands having odd indices. This inflection of H and H′ is illustrated in Fig. 9 for the imaginary parts of the subbands I_{k+2} to I_{k-2}. In addition, the index "m" stands for "mid" and refers to the contribution of the center subband signal. Furthermore, the index "l" stands for "low" and considers the contribution of the lower subband shown in Fig. 8to the current subband, i.e. the subband having an index lower by 1. In analogy, "u" stand for "up" and refers to the contribution of the subband illustrated at the top in Fig. 8for the current subband, i.e. to the subbands having an index higher by 1.

The synthesis filters G corresponding to the individual analysis filters H are illustrated in Fig. 8. G_l has a high-pass characteristic, whereas G_u has a low-pass characteristic. Thus, as has been described before, the product of $G_{u'}$ and H_u is the same as the product of H_l and H_u or the product of $G_{l'}$ and H_l is the same as the product of H_u and H_l and nearly equals 0, since here a respective high-pass filter is multiplied by a low-pass filter and the resulting frequency response of a high-pass filter and a low-pass filter having similar cutoff frequencies equals 0 and/or approximates 0. Even for cases in which the cutoff frequencies are not identical but are apart, the resulting frequency response equals 0. If the cutoff frequency of the low-pass filter is smaller than the cutoff frequency of the high-pass filter, the resulting frequency response will also equal 0. Only in the case in which the cutoff frequency of the low-pass filter is higher than the cutoff frequency of the high-pass filter, the approximation given above would not hold true. However, such a situation does not occur in typical polyphase filter banks and/or would, if occurring, only result in slight interferences which would result in a somewhat more imprecise correction term. Due to the fact that the correction term is preferably weighted by the difference of the two weighting factors

concerned, this error would also decrease with a decreasing difference.

Fig. 3 shows a schematic illustration of the preferred filters derived above which are implemented by the inventive correction term determiner 12. It becomes obvious from Fig. 3a that the entire device includes a filter part 30 and a weighting part 31. The weighter 10 of Fig. 1 symbolized in the weighting part 31 of Fig. 3a by c_k is in the weighting part 31. The combiner 13 of Fig. 1 corresponds to the adder 13 in Fig. 3a. The correction term determiner 12 includes the filter actions with the four filters H_{lm}, H_{ll}, H_{uu} and H_{um}. In addition, the correction term determiner also includes the weighting of the unweighted correction terms L_k and U_k by the difference of the respective two weighting factors concerned, i.e. by q_k and p_k, respectively, as is indicated in the weighting part 31. A more detailed implementation of the filter part of Fig. 3a is illustrated in Fig. 3b. The subband signal x_{k-1} is fed to the low-pass filter H_{lm} 32. In addition, the subband signal x_k is fed to the low-pass filter H_{ll} 33. In addition, the subband signal x_k is fed to the high-pass filter H_{uu} 34 and in addition the next subband signal x_{k+1} is fed to the filter H_{um} 35 which may also be implemented as a high-pass filter. The output signals of filters 32 and 33 are combined in an adder 34 and represent a first unweighted correction term l_k. Additionally, the output signals of the filters 34 and 35 are added in an adder 35 and represent a second unweighted correction term u_k. In addition, the delay of the filters occurring when the filters are implemented as digital filters, i.e. FIR or IIR filters, is considered for the subband signal x_k which is weighted by the weighting factor c_k provided for this subband signal. This considering of delays of the filters 33 to 35 takes place in a delay stage 38 and can occur before or after weighting. In order for such an implementation to achieve maximum quality, it is preferred for all the filter lengths 32, 33, 34, 35 to be identical

and for the delay 38 to be adjusted to the filter length of the filters 32 to 35. If, for example, the filters 32 to 35 each have a filter length of 11, the delay 38 has to provide a delay magnitude of five subband signal samples.

While Figs. 3a and 3b illustrate the situation in which the filters 32, 33, 34, 35 are represented as product filters, i.e. as filters for calculating the terms l_k, u_k which then only have to be weighted, Figs. 3c and 3d show an embodiment of an implementation of the present invention where the correction term is not calculated by 4 product filters, but all in all 6 individual filters 320, 330, 340, 350, 381, 382.

As is illustrated in particular in Fig. 3c, the signal L_k is calculated by filtering X_{k-1} by the filter H_m and by adding the filtered signal X_k having been filtered by H_l. Again, the normalization factor of 0.5 has been introduced. This normalization factor, however, can be omitted, as is the case in the first embodiment, or be set to a different value, including 1. Furthermore, the other component U_k is calculated by filtering X_k by H_u, wherein $X_{k+1} \cdot H_m$ is subtracted from $X_k \cdot H_u$. In contrast to the equations shown under Fig. 3a, where the products have already been considered in the filters, the signals in Fig. 3c are filtered individually. The results L_k and U_k will then, as is shown in Fig. 3a, be weighted by p_k and q_h, respectively. In addition to this weighting, filtering by H_l and H_u is performed.

In contrast to Fig. 3a, there is a first filter part and additionally a second filter part which may be integrated and/or combined with the weighting part. The weighting factors can thus already be considered in the filter coefficients or can be applied separately before or after filtering by the digital filter H_l and/or H_u. Thus, the delays z^{-d} consider the delay caused by the filtering in the first filter part of the two components X_{k-1} and/or X_{k+1}

and additionally consider the delays in the second filter part caused by the filtering of L_k and/or U_k which are filtered by the filters H_l and/or H_u.

Although, depending on the filter bank implemented, any filter characteristics can be used for the filters H_n, H_l, H_u, it is preferred to use a low-pass filter for H_l, it is preferred to use a high-pass filter for H_u and/or it is also preferred to use a bandpass filter for H_m. The filter H_l has a similar form as Fig. 10, since H_{ll} 100 in Fig. 10 equals the square of the filter H_l. The filter H_u implemented as a high-pass filter results by mirroring the left part in Fig. 10 at a vertical axis at $\pi/2$, i.e. about in the center of Fig. 10. The filter H_{lm} which is no longer there in Fig. 3c since it is a product filter of a bandpass filter and a low-pass filter, could be mirrored at the line at $\pi/2$ to obtain the filter H_{um} 35 in Fig. 3b, although this product filter in an assembled form is no longer there in Fig. 3c, but is first calculated implicitly before then combining the components by the combiner 13.

Whereas in Fig. 3b the correction term determiner 12 of Fig. 3a is implemented by the filter part 30 and the weighting of the components L_k, U_k by the weighting factors p_k and q_k, the correction term determination according to Figs. 3c and 3d takes place in a kind of double filter stage, wherein at first the signals L_k, U_k at the output of the summers 360 and/or 370 are calculated, not using product filters, but using the individual filters, wherein in the second filter part the weighting by p_k and/or q_k is then performed with subsequent individual filtering.

The weighting of the subband signal X_k by the weighter 10, however, takes place in Fig. 3d like in Fig. 3a.

In the embodiment shown in Fig. 3c and Fig. 3d or put generally, two filters are not united to form a product filter. Instead, they are implemented as individual

filters. Even if there is no uniting in the product filter, there is - apart from the implementation - still the advantage of shortened filter lengths. Thus, the delay compared to a direct recalculation from real to complex and/or complex to real is reduced. The swung dash above the filters in blocks 320, 330, 340, 350, 381, 382 means that the filters, as are schematically indicated in Fig. 10 for the product filters, are reduced in their filter lengths compared to a subband filter of a normal filter bank. It is preferred to use filter lengths which are smaller than a filter length of a subband filter to generate the subband signals x_{k-1}, x_k and/or x_{k+1}. In addition, it is preferred for the filter lengths of the filters h_u, h_m, h_l after the approximation, i.e. after shortening, to be, like in the other case, at most 50% of the length of a filter which has been used to generate a subband signal by applying several such filters in a subband filter bank.

Preferably, filter lengths of < 21 are preferred, wherein the delay of such filters is < 10. The implementation shown in Fig. 3d provides, in comparison to the implementation shown in Figs. 3a and b, advantages with quickly time-variable attenuation factors. With regard to the time form, the implementation shown in Fig. 3d is more similar to the real/complex-complex/real implementation, whereas in the product filter realization, no more filtering takes place after applying the amplification factors.

Irrespective of whether a realization with individual shortened filters or a summary with product filters is chosen, aliasing-reduced quick real filter banks are implemented according to the invention. In specially preferred embodiments, the filter lengths in Fig. 3d are even reduced compared to the filter lengths in Fig. 3b in that the entire calculation in Fig. 3d has a similar delay as the entire calculation in Fig. 3b. An implementation similar to Fig. 3b would then in Fig. 3d be for the filters in the first filter part to have a filter length of 7

coefficients, which would correspond to a delay magnitude of 3 samples of a subband signal. In this case, the second delay 383 and/or the subsequent filters 381, 382, for example, would have a filter length of 4 to implement a delay of 2. It is pointed out here that somewhat longer or somewhat shorter filters and/or an implementation in Fig. 3d will also bring about advantages when the overall delay is somewhat greater than the product filter delay of Fig. 3b.

Fig. 4 shows the usage of the device for weighting described in Figs. 1 to 3d in an analysis filter band and/or synthesis filter bank. It becomes obvious from Fig. 4 that for every filter channel 0 to N-1, one device as shown in Fig. 1 is necessary. Preferably, however, every device for processing has, when implemented like in Fig. 3b, the same four filters 32 to 35 so that only the same four filters have to be calculated and/or optimized irrespective of the number of subband signals and/or filter channels of an analysis/synthesis filter bank.

The actual calculation of the filters may be performed either by a direct calculation from the analysis/synthesis prototype filters or by a numerical optimization which typically takes place in a computer-aided manner. In such a numerical optimization of filters 32 to 35, a filter length is preset so that a set of filters for different filter lengths can be obtained. As is in particular illustrated in Fig. 10, the marked low-pass characteristic is obtained for the filter with the filter transfer function 100, i.e. H_{11}, or for the filter with the filter transfer function 101, i.e. H_{1m}. However, it can be seen that these filter with a very marked attenuation in the blocking region can be approximated by considerably shorter filters, namely by the filters 102 or 103. The filters 102 and 103 have a filter length of only 11 and thus approximate the filters 100 and 101. However, it can be seen that in the region of low frequencies, the deviations are very low and only increase

at higher frequencies. However, on the other hand, a blocking attenuation of greater than 40 dB is ensured by the filters 102 and 103 so that these very short filters already cause good aliasing suppression.

Fig. 11 shows a response of the filter bank for an impulse at a position 8 in a subband sample period. A real filter bank provides a form indicated at 110. A complex filter bank provides a form indicated at 112. A real filter bank including correction according to the present invention provides the form indicated at 111. It can be seen that the real filter bank including correction has nearly the same form as the complex filter bank, however can be implemented considerably cheaper. Only directly at the band limit between k-1 and k does the real filter bank including correction exhibit a wavy form, which can be attributed to the fact that only filter lengths of 11, as is illustrated in Fig. 10, instead of complete filters, as were illustrated in Fig. 10, are used. It is apparent that the deviation between the real filter bank including correction and the complex filter bank which is aliasing-resistant is negligible although short filters 32 to 35 of Fig. 3b are already used. The deviation between the forms 111 and 112 becomes greater with shorter filter lengths, however, for a delay-optimized variation, filter lengths smaller than 5 may be used, wherein the deviation between the curves 111 and 112 still remains reasonable.

Subsequently, a filter bank response when a sine tone at 1% above the band limit is considered is shown in Fig. 12. The input signal 121 represents the sine tone. A real filter bank would produce aliasing, as is illustrated by a curve 122. Aliasing becomes noticeable by the "secondary peak" 125, wherein this secondary peak is caused by the fact that the neighboring bands k-1 and k have been weighted by different weighting factors. Again, it is apparent that a complex filter bank does not have such a secondary peak, i.e. does not generate such aliasing, and that the complex

filter bank is approximated optimally by a real filter bank including correction, wherein a deviation of the real filter bank to the complex filter bank is only in the region 126. The real filter bank provides a greater attenuation than the complex filter bank, which in turn can be attributed to the fact that the filter lengths of the filters 32 to 35 have been shortened to 11.

It is to be pointed out that in the example in Fig. 12, and in the examples in Figs. 13, 14, 15, 16, 17, there has always been an attenuation of a subband by 20 dB compared to the other subband.

Fig. 13 shows a similar case as Fig. 12, however with a sine tone at 5% above the band limit. Again, a real filter bank would generate the secondary peak 125. This secondary peak, however, is attenuated nearly completely by the real filter bank including correction 124. Only a very small deviation at 127 is still to be seen. When reducing the filter length of the filters 32 to 35, this peak 127 would continue increasing. Even with degenerated filters, i.e. filters only performing weighting by a weighting factor, the peak 127 would still be smaller than the secondary peak 125. By inventively weighting the filtered values by the difference of the two weighting factors, however, at least for the case of identical or nearly identical weighting factors, almost no interference at all would be introduced, despite the rather rudimentary filtering by degenerated filters.

Fig. 14, Fig. 15, Fig. 16 and Fig. 17 show similar scenarios, wherein the sine tone, however, is in an ever greater distance to the band limit. All the pictures clearly show the aliasing component which would be generated by a real filter bank if no inventive correction was performed. In addition, all the pictures show even smaller deviations at 127 between the real filter bank

including correction according to the present invention and an aliasing-resistant complex filter bank 123.

Fig. 18 shows a scenario similar to that of Figs. 12 to 17, however for a transform in the form of the MDCT. Again, the clear aliasing component 125 can be recognized, occurring at a frequency of 127.88. By means of the inventive correction, i.e. by filtering corresponding MDCT coefficients of successive MDCT peaks by the filters 32 to 35, the aliasing component is reduced except for the small deviation at 127 in Fig. 18. If filter lengths of 11 are used for the filters 32 to 35, the entire assembly will only be settled after around 10 MDCT spectra. Thus, in the implementation of Figs. 3a and 3b, a delay of 5 MDCT blocks is necessary until sensitive output values are generated.

Depending on the circumstances, the inventive method may be implemented in either hardware or software. The implementation can be on a digital storage medium, in particular on a disc or a CD having control signals which may be read out electronically, which can cooperate with a programmable computer system such that the corresponding method will be performed. Generally, the invention is thus also in a computer program product having a program code stored on a machine-readable carrier for performing the inventive method when the computer program product runs on a computer. Put differently, the invention may thus also be realized as a computer program having a program code for performing the method when the computer program runs on a computer.

Claims

1. A device for processing a real subband signal (x_k) of a plurality of real subband signals which are a representation of a real discrete-time signal x(n) generated by an analysis filter bank (50), comprising:

a weighter (10) for weighting the subband signal (x_k) by a weighting factor (c_k) determined for the subband signal to obtain a weighted subband signal (11);

a correction term determiner (12) for calculating a correction term, the correction term determiner being implemented to calculate the correction term using at least one other subband signal (x_l) and using another weighting factor (c_l) provided for the other subband signal (x_l), the other weighting factor (c_l) differing from the weighting factor (c_k); and

a combiner (13) for combining the weighted subband signal and the correction term to obtain a corrected subband signal (y_k).

2. The device according to claim 1, wherein the correction term determiner (12) is implemented to generate the correction term in dependence on a difference of the weighting factor (c_k) of the subband signal and the other weighting factors (c_l) for the further subband signal (x_l).

3. The device according to claim 1 or 2, wherein the correction term determiner (12) is implemented to determine the correction term in dependence on the subband signal (x_k).

4. The device according to one of the preceding claims, wherein the correction term determiner (12) is implemented to use as the other subband signal a

subband signal (x_{k+1}, x_{k-1}) of the plurality of subband signals having a frequency range index differing from a frequency range index of the subband signal (x_k) by "1".

5. The device according to one of the preceding claims, wherein the correction term determiner (12) is implemented to determine another correction term (12b) which is dependent on a third subband signal (x_{k-1}) and a third weighting factor (c_{k-1}) associated to the third subband signal (x_{k-1}), the third weighting factor (c_{k-1}) differing from the weighting factor (c_k).

6. The device according to claim 5, wherein the combiner (13) is implemented to combine the other correction term and the weighted subband signal (11).

7. The device according to claim 5 or 6, wherein the correction term determiner (12) is implemented to use as a third subband signal a subband signal having a frequency range index differing from a frequency range index of the subband signal (x_k) and from a frequency range index (k+1) of the other subband signal (x_{k+1}).

8. The device according to one of the preceding claims, wherein the correction term determiner is implemented to calculate the correction term as a linear combination of the subband signal filtered by a first filter (33) and the other subband signal (x_{k-1}) filtered by another filter (32), the filters (32, 33) being low-pass filters or high-pass filters having a filter length between 1 and 20.

9. The device according to one of the preceding claims, wherein the correction term determiner is implemented to calculate an unweighted correction term in dependence on the other subband signal (x_{k-1}), and to weight the unfiltered correction term by a difference

of the weighting factor (c_k) for the subband (x_k) and the weighting factor (c_l) for the subband (x_l).

10. The device according to one of the preceding claims,

wherein the correction term determiner (12) is implemented to calculate the correction term (KT) according to the following equation:

$KT = p_k L_k(z) + q_k U_k(z)$,

p_k equaling a difference between the weighting factor c_{k-1} for a neighboring subband signal x_{k-1} and the weighting factor c_k of the subband signal x_k, q_k being a difference between another neighboring subband signal x_{k+1} and the weighting factor c_k of the subband signal x_k, L_k being an unweighted correction term of the one neighboring subband signal x_{k-1}, and U_k being an unweighted correction term caused by another neighboring subband signal x_{k+1}.

11. The device according to claim 10, wherein the unweighted correction terms L_k and U_k are calculated as follows:

$$L_k(z) = A_1\left(H_{ll}(z)X_k(z) - H_{lm}(z)X_{k-1}(z)\right)$$
$$U_k(z) = A_2\left(H_{uu}(z)X_k(z) - H_{um}(z)X_{k+1}(z)\right)$$

A_1 and A_2 being factors, H_{ll} representing a first low-pass filter (33), H_{lm} representing a second low-pass filter, H_{uu} representing a first high-pass filter (34), and H_{um} representing a second high-pass filter (35).

12. The device according to claim 11, wherein a cutoff frequency of the high-pass filters (34, 35) is greater

than or equal to a cutoff frequency of the low-pass filters (32, 33).

13. The device according to one of the preceding claims,

wherein the analysis filter bank may be implemented by a transform of a block of samples to a spectral representation, and a subband signal comprises subband samples comprising spectral coefficients of the same frequency index from a sequence of successive spectral representations.

14. The device according to one of claims 1 to 12, wherein the analysis filter bank (50) is a decimated filter bank comprising filters describable by modulation of a prototype filter.

15. The device according to claim 13 or 14, wherein a subband signal is a signal having several samples, N/M samples being generated for each signal from a quantity of N values of the discrete-time signal, M being a number of subband signals generated by the analysis filter bank.

16. The device according to one of the preceding claims, wherein the combiner is implemented to perform an addition of the weighted subband signal and the correction term.

17. The device according to one of the preceding claims, further comprising:

means for providing the different weighting factors associated to the subband signals, the means (55) for providing being implemented to determine the several weighting factors due to an equalizer function or echo suppression or bandwidth extension or parametric multi-channel encoding.

18. The device according to claim 10, wherein the unweighted correction terms L_k and U_k are calculated using the following equations:

$$L_k' = (H_l \cdot X_k - H_m \cdot X_{k-1}) \cdot A_1$$

$$U_k' = (X_k \cdot H_u - H_m \cdot X_{k+1}) \cdot A_2$$

A_1, A_2 being constant factors, H_m, H_l, H_u representing filters, the filter H_l (330) comprising a low-pass characteristic, and the filter H_u (340) comprising a high-pass characteristic.

19. The device according to claim 18, wherein the filter H_m (320, 350) has a bandpass characteristic.

20. The device according to claim 18 or 19, wherein the unweighted correction terms L_k and U_k are calculated using the following equations:

$$L_k = L_k' \cdot H_l$$
$$U_k = U_k' \cdot H_u$$

H_l being a low-pass filter (381), H_u being a high-pass filter (382) and L_k' and U_k' being intermediate signals.

21. The device according to one of the preceding claims, wherein the correction term determiner (12) comprises filters (32, 33, 34, 35; 320, 330, 340, 350, 381, 383, 382) comprising a filter length which is smaller than a filter length of a subband filter to generate the subband signal (x_k) or the other subband signal (x_l).

22. The device according to claim 21,

wherein the correction term determiner (12) comprises filters (32, 33, 34, 35; 320, 330, 340, 350, 381, 383, 382) having a filter length which is selected such that a delay of a filter is smaller than 10 subband signal samples of the subband signal (x_k) or the other subband signal (x_l).

23. The device according to claim 22,

wherein the correction term determiner (12) only comprises filters the filter length of which is selected such that a delay of every filter is smaller than 6 values of the subband signal (x_k) or the other subband signal (x_l).

24. A method for processing a real subband signal (x_k) of a plurality of real subband signals which are a representation of a real discrete-time signal x(n) generated by an analysis filter bank (50), comprising the steps of:

weighting (10) the subband signal (x_k) by a weighting factor (c_k) determined for the subband signal to obtain a weighted subband signal (11);

calculating (12) a correction term using at least one other subband signal (x_l) and using another weighting factor (c_l) provided for the other subband signal (x_l), the other weighting factor (c_l) differing from the weighting factor (c_k); and

combining (13) the weighted subband signal and the correction term to obtain a corrected subband signal (y_k).

25. An analysis filter bank comprising:

means (50) for generating a plurality of subband signals from a discrete-time signal; and

a device for processing according to one of claims 1 to 23 for each subband signal to obtain processed subband signals.

26. A synthesis filter bank comprising:

for every subband signal of a plurality of subband signals having been generated by an analysis filter bank, a device according to one of claims 1 to 23 to obtain processed subband signals; and

a plurality of synthesis filters (51) for filtering the processed subband signals to obtain synthesis-filtered subband signals; and

a summer (58) for summing the filtered subband signals to obtain a discrete-time signal.

27. A method for filtering a discrete-time signal, comprising the steps of:

generating (50) a plurality of subband signals from the discrete-time signal; and

for every subband signal, processing the subband signal by a method according to claim 24 to obtain processed subband signals.

28. A method for synthesizing a signal, comprising the steps of:

for every subband signal of a plurality of subband signals having been generated by an analysis filter bank, processing the subband signal by a method

according to claim 24 to obtain processed subband signals; and

synthesis-filtering (51) the processed subband signals to obtain synthesis-filtered subband signals; and

summing (58) the filtered subband signals to obtain a synthesized signal.

29. A computer program comprising a program code for performing the method according to claim 24, 27 or 28 when the computer program runs on a computer.