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(54) **METHOD AND DEVICE FOR ADAPTIVE BANDWIDTH PITCH SEARCH IN CODING WIDEBAND SIGNALS**

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G10L 19/00 (2006.01)
G10L 19/12 (2006.01)

(52) **U.S. Cl.** **704/207**; 704/219; 704/223

(58) **Field of Classification Search** 704/207, 704/219, 220, 223

See application file for complete search history.

(56) **References Cited**

U.S. PATENT DOCUMENTS

5,444,816	A	8/1995	Adoul et al.	
5,621,852	A	4/1997	Lin	
5,699,482	A	12/1997	Adoul et al.	
5,701,392	A	12/1997	Adoul et al.	
5,754,976	A *	5/1998	Adoul et al.	704/223
7,260,521	B1 *	8/2007	Bessette et al.	704/219
7,672,837	B2 *	3/2010	Bessette et al.	704/207

FOREIGN PATENT DOCUMENTS

EP	0331858	A	9/1989
EP	0421444	A	4/1991
EP	0722165	A	7/1996

OTHER PUBLICATIONS

M. V. Nazarov et al., "Techniques of Digital Processing and Transmission of Voice Signals"; *Moscow, Radio i Svyaz' Publishers*, pp. 10-12 (1985).

Bishnus S. Atal et al., "Predictive Coding of Speech Signals and Subjective Rrror Criteria" IEEE Transaction ASSP, vol. 27, No. 3, pp. 247-254 (1979).

* cited by examiner

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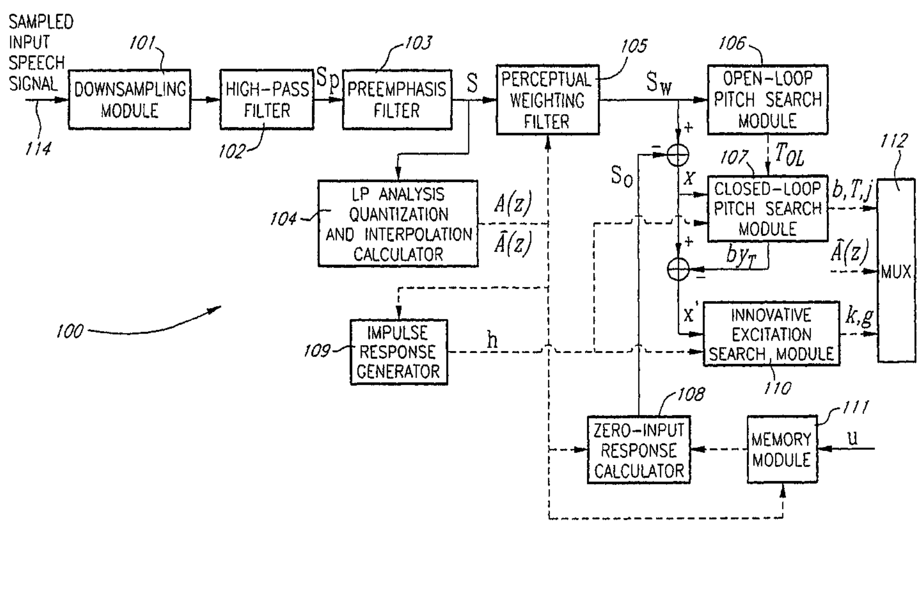
(57) **ABSTRACT**

A pitch search method and device for digitally encoding a wideband signal, in particular but not exclusively a speech signal, in view of transmitting, or storing, and synthesizing this wideband sound signal. The new method and device which achieve efficient modeling of the harmonic structure of the speech spectrum uses several forms of low pass filters applied to a pitch codevector, the one yielding higher prediction gain (i.e. the lowest pitch prediction error) is selected and the associated pitch codebook parameters are forwarded.

18 Claims, 4 Drawing Sheets



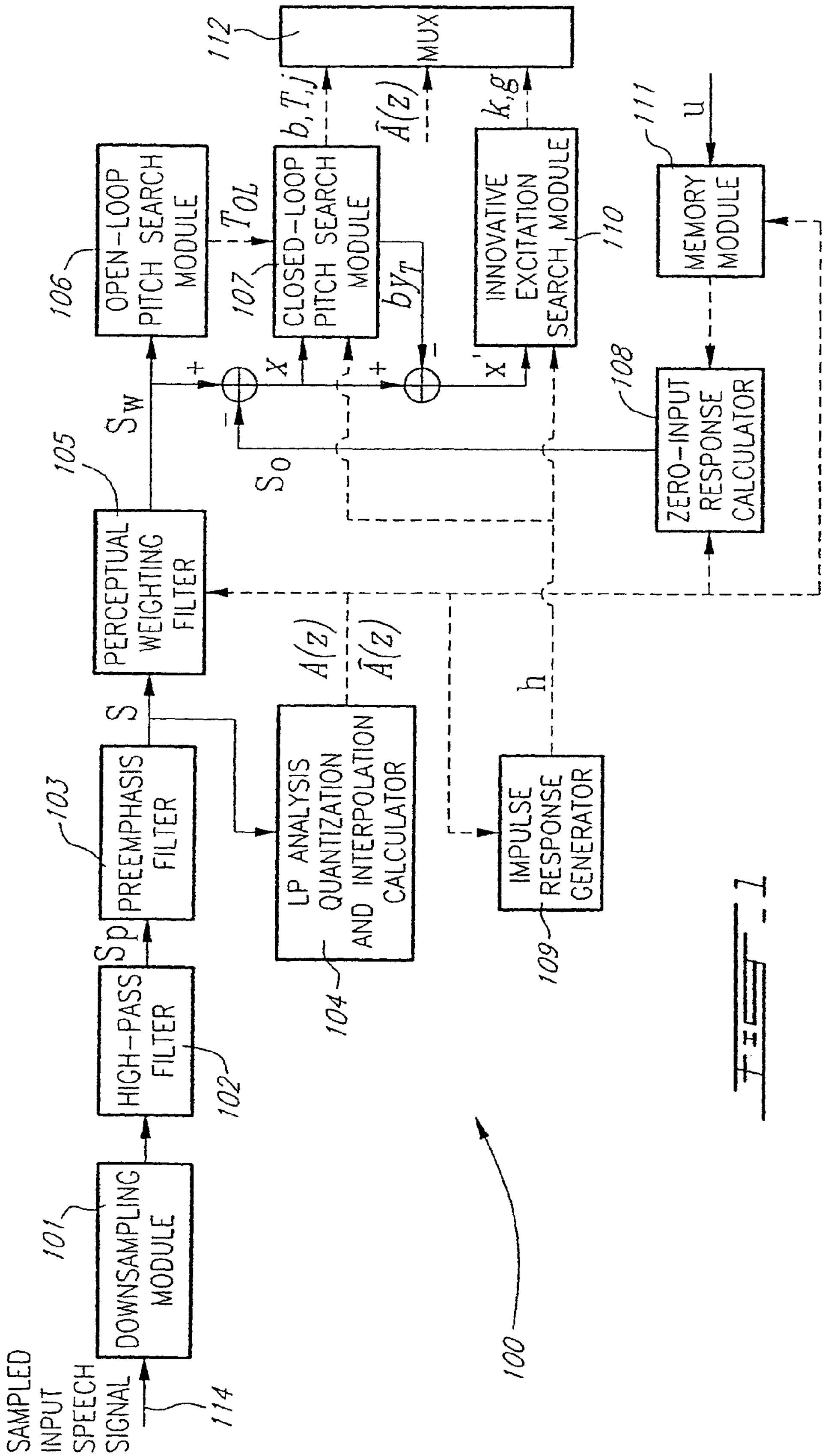

SAMPLED INPUT SPEECH SIGNAL
114
DOWNSAMPLING MODULE
101
HIGH-PASS FILTER
102
Sp
PREEMPHASIS FILTER
103
S
PERCEPTUAL WEIGHTING FILTER
105
Sw
LP ANALYSIS QUANTIZATION AND INTERPOLATION CALCULATOR
104
A(z)
Â(z)
IMPULSE RESPONSE GENERATOR
109
h
OPEN-LOOP PITCH SEARCH MODULE
106
T_OL
CLOSED-LOOP PITCH SEARCH MODULE
107
b,T,j
by_T
x
x'
So
INNOVATIVE EXCITATION SEARCH MODULE
110
k,g
Â(z)
MUX
112
ZERO-INPUT RESPONSE CALCULATOR
108
MEMORY MODULE
111
u
100

FIG. 1

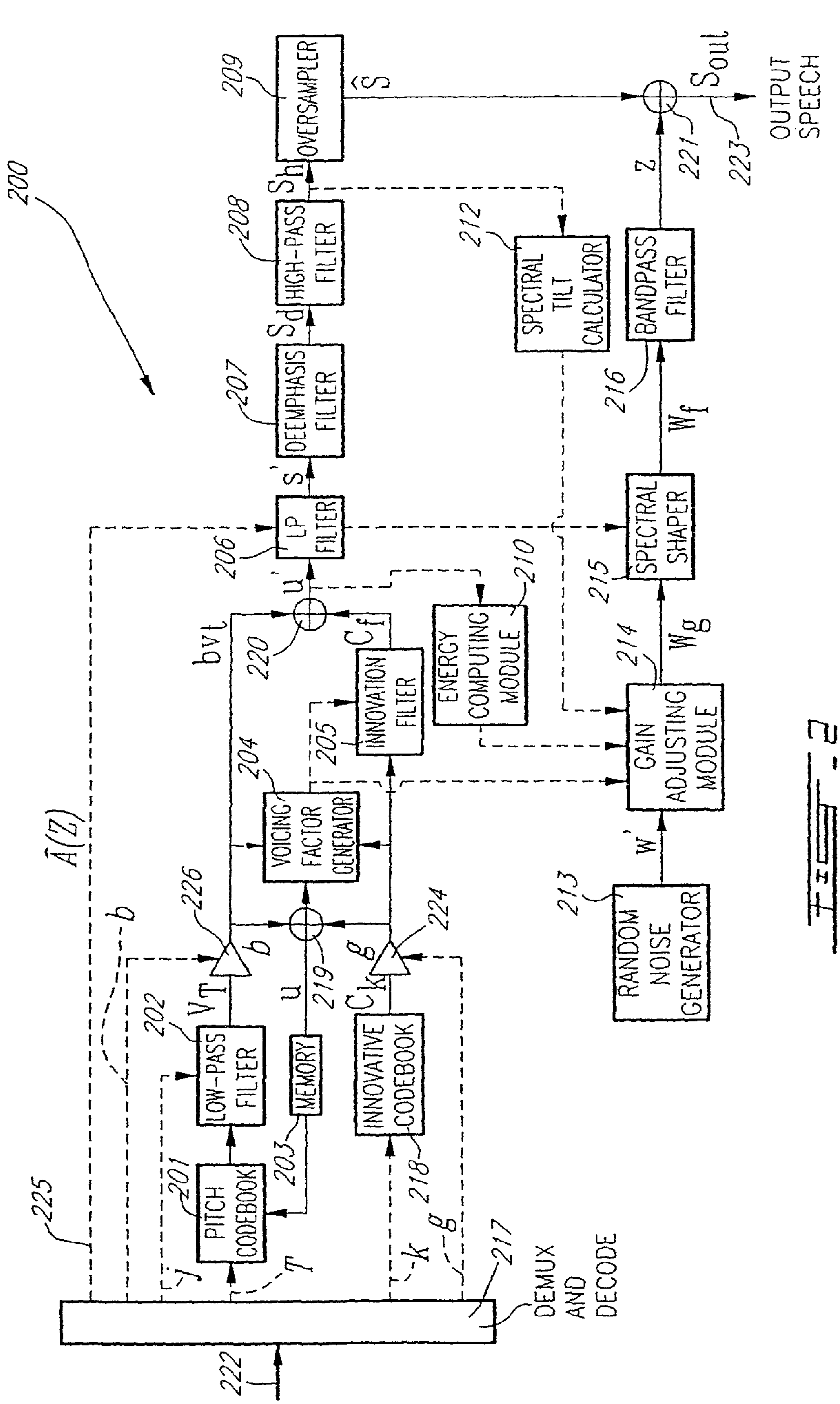

200
PITCH CODEBOOK
LOW-PASS FILTER
MEMORY
INNOVATIVE CODEBOOK
VOICING FACTOR GENERATOR
INNOVATION FILTER
LP FILTER
DEEMPHASIS FILTER
HIGH-PASS FILTER
OVERSAMPLER
ENERGY COMPUTING MODULE
SPECTRAL TILT CALCULATOR
RANDOM NOISE GENERATOR
GAIN ADJUSTING MODULE
SPECTRAL SHAPER
BANDPASS FILTER
DEMUX AND DECODE
OUTPUT SPEECH
201
202
203
204
205
206
207
208
209
210
212
213
214
215
216
217
218
219
220
221
222
223
224
225
226

FIG. 2

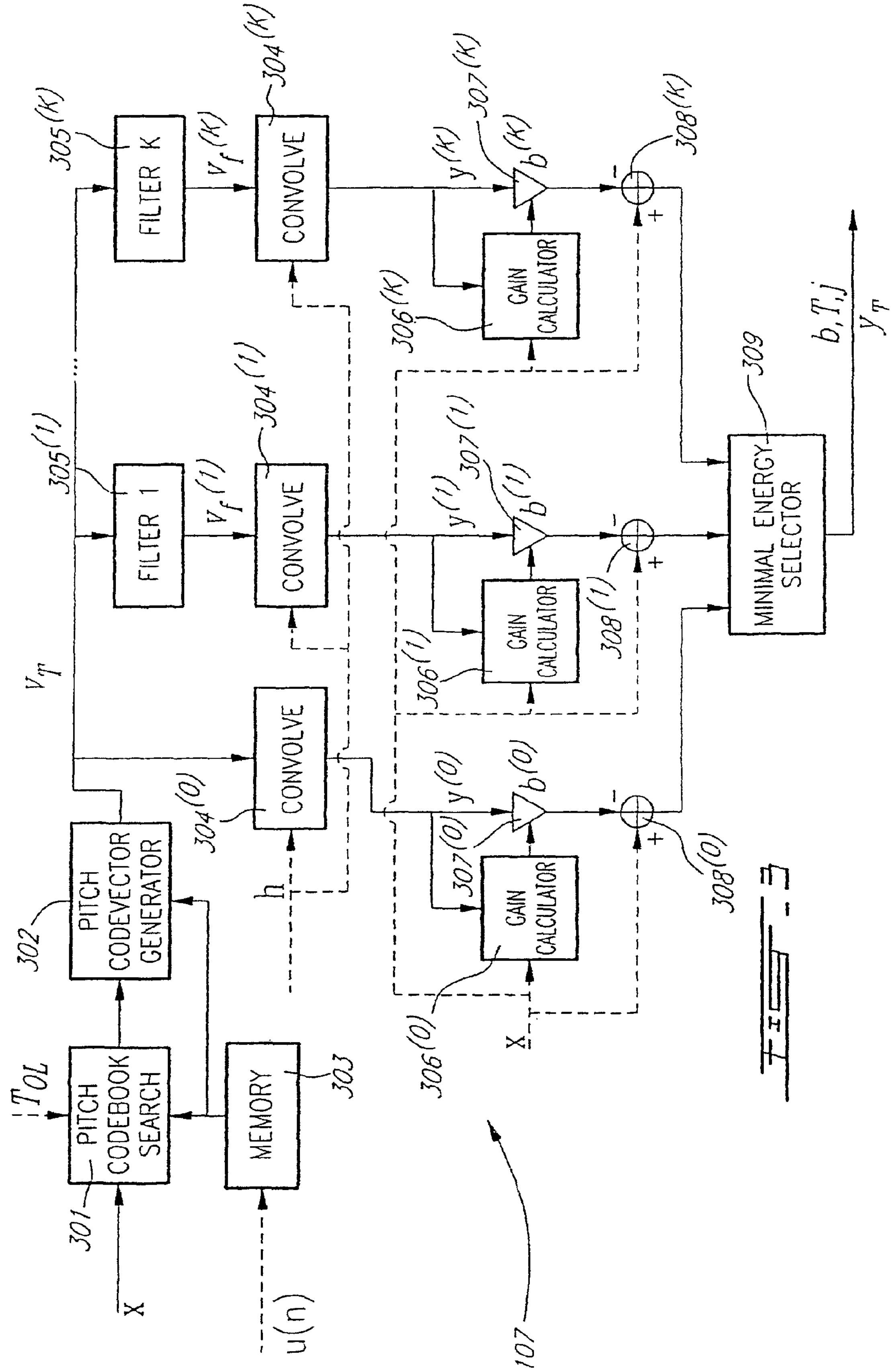

PITCH CODEBOOK SEARCH
PITCH CODEVECTOR GENERATOR
MEMORY
FILTER 1
FILTER K
CONVOLVE
GAIN CALCULATOR
MINIMAL ENERGY SELECTOR
X
u(n)
h
T_{OL}
v_T
$v_f^{(1)}$
$v_f^{(K)}$
$y^{(0)}$
$y^{(1)}$
$y^{(K)}$
$b^{(0)}$
$b^{(1)}$
$b^{(K)}$
b, T, j
y_T
301
302
303
$304^{(0)}$
$304^{(1)}$
$304^{(K)}$
$305^{(1)}$
$305^{(K)}$
$306^{(0)}$
$306^{(1)}$
$306^{(K)}$
$307^{(0)}$
$307^{(1)}$
$307^{(K)}$
$308^{(0)}$
$308^{(1)}$
$308^{(K)}$
309
107

FIG. 3

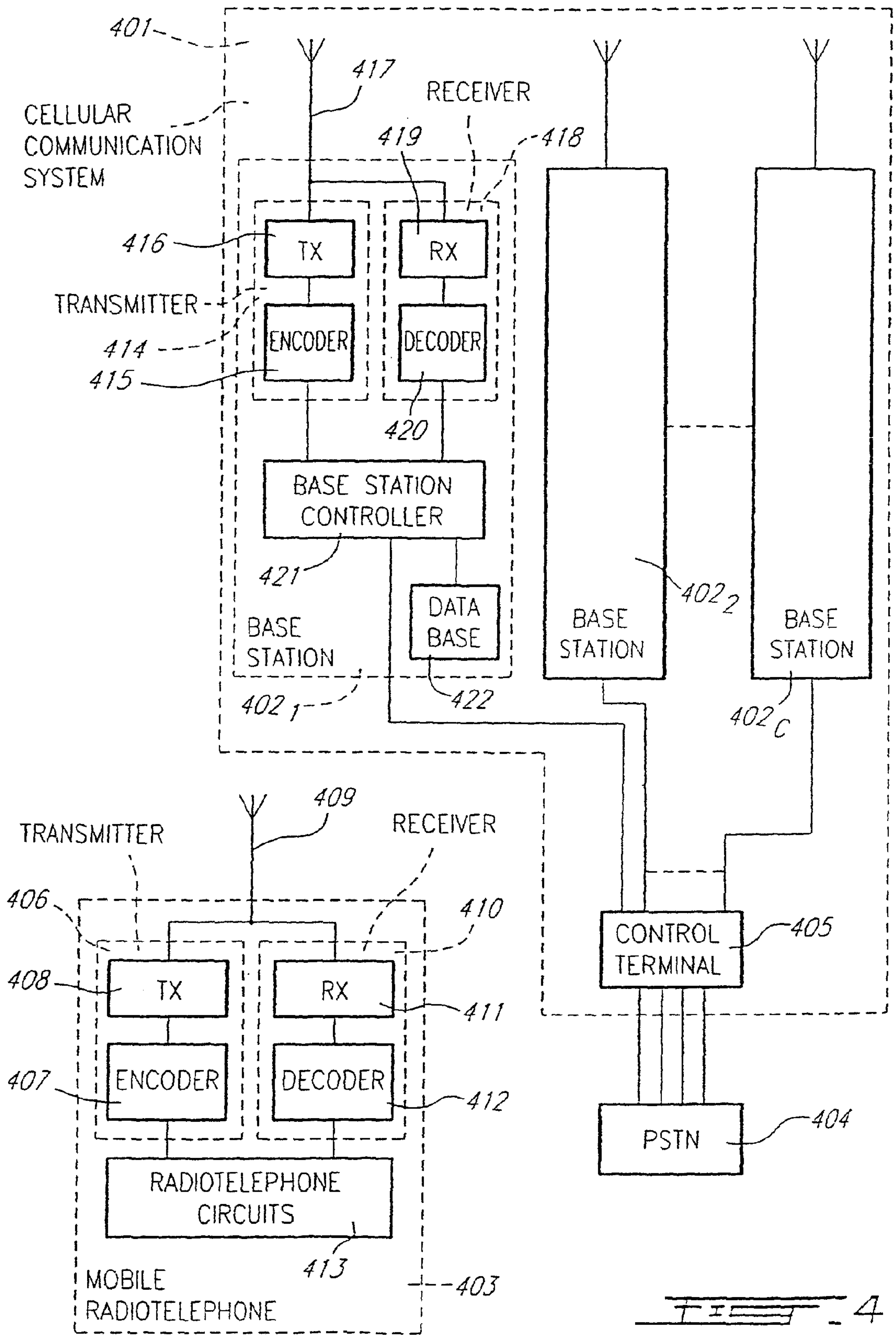

401
CELLULAR COMMUNICATION SYSTEM
417
RECEIVER
419
418
416
TX
RX
TRANSMITTER
414
ENCODER
DECODER
415
420
BASE STATION CONTROLLER
421
DATA BASE
BASE STATION
402 1
422
402 2
BASE STATION
BASE STATION
402 C
409
TRANSMITTER
RECEIVER
406
410
408
TX
RX
411
407
ENCODER
DECODER
412
CONTROL TERMINAL
405
PSTN
404
RADIOTELEPHONE CIRCUITS
413
MOBILE RADIOTELEPHONE
403

FIG. 4

METHOD AND DEVICE FOR ADAPTIVE BANDWIDTH PITCH SEARCH IN CODING WIDEBAND SIGNALS

This patent application is a continuation application of Ser. No. 11/498,771, filed Aug. 4, 2006, now U.S. Pat. No. 7,672,837 which is a divisional application of Ser. No. 09/830,114, filed Jun. 20, 2001 now U.S. Pat. No. 7,260,521 which is the national phase under 35 U.S.C. §371 of PCT International Application No. PCT/CA99/01008 which has an International filing date of Oct. 27, 1999, which designated the United States of America and was published in English, the entire contents of all of which are hereby incorporated by reference in their entirety.

BACKGROUND OF THE INVENTION

1. Field of the Invention

The present invention relates to an efficient technique for digitally encoding a wideband signal, in particular but not exclusively a speech signal, in view of transmitting, or storing, and synthesizing this wideband sound signal. More specifically, this invention deals with an improved pitch search device and method.

2. Brief Description of the Prior Art

The demand for efficient digital wideband speech/audio encoding techniques with a good subjective quality/bit rate trade-off is increasing for numerous applications such as audio/video teleconferencing, multimedia, and wireless applications, as well as Internet and packet network applications. Until recently, telephone bandwidths filtered in the range 200-3400 Hz were mainly used in speech coding applications. However, there is an increasing demand for wideband speech applications in order to increase the intelligibility and naturalness of the speech signals. A bandwidth in the range 50-7000 Hz was found sufficient for delivering a face-to-face speech quality. For audio signals, this range gives an acceptable audio quality, but still lower than the CD quality which operates on the range 20-20000 Hz.

A speech encoder converts a speech signal into a digital bitstream which is transmitted over a communication channel (or stored in a storage medium). The speech signal is digitized (sampled and quantized with usually 16-bits per sample) and the speech encoder has the role of representing these digital samples with a smaller number of bits while maintaining a good subjective speech quality. The speech decoder or synthesizer operates on the transmitted or stored bit stream and converts it back to a sound signal.

One of the best prior art techniques capable of achieving a good quality/bit rate trade-off is the so-called Code Excited Linear Prediction (CELP) technique. According to this technique, the sampled speech signal is processed in successive blocks of L samples usually called frames where L is some predetermined number (corresponding to 10-30 ms of speech). In CELP, a linear prediction (LP) filter is computed and transmitted every frame. The L-sample frame is then divided into smaller blocks called subframes of size N samples, where L=kN (N and k is the number of subframes in a frame (N usually corresponds to 4-10 ms of speech). An excitation signal is determined in each subframe, which usually consists of two components: one from the past excitation (also called pitch contribution or adaptive codebook) and the other from an innovation codebook (also called fixed codebook). This excitation signal is transmitted and used at the decoder as the input of the LP synthesis filter in order to obtain the synthesized speech.

An innovation codebook in the CELP context, is an indexed set of N-sample-long sequences which will be referred to as N-dimensional codevectors. Each codebook sequence is indexed by an integer k ranging from 1 to M where M represents the size of the codebook often expressed as a number of bits b, where $M=2^b$.

To synthesize speech according to the CELP technique, each block of N samples is synthesized by filtering an appropriate codevector from a codebook through time varying filters modeling the spectral characteristics of the speech signal. At the encoder end, the synthetic output is computed for all, or a subset, of the codevectors from the codebook (codebook search). The retained codevector is the one producing the synthetic output closest to the original speech signal according to a perceptually weighted distortion measure. This perceptual weighting is performed using a so-called perceptual weighting filter, which is usually derived from the LP filter.

The CELP model has been very successful in encoding telephone band sound signals, and several CELP-based standards exist in a wide range of applications, especially in digital cellular applications. In the telephone band, the sound signal is band-limited to 200-3400 Hz and sampled at 8000 samples/sec. In wideband speech/audio applications, the sound signal is band-limited to 50-7000 Hz and sampled at 16000 samples/sec.

Some difficulties arise when applying the telephone-band optimized CELP model to wideband signals, and additional features need to be added to the model in order to obtain high quality wideband signals. Wideband signals exhibit a much wider dynamic range compared to telephone-band signals, which results in precision problems when a fixed-point implementation of the algorithm is required (which is essential in wireless applications). Further, the CELP model will often spend most of its encoding bits on the low-frequency region, which usually has higher energy contents, resulting in a low-pass output signal. To overcome this problem, the perceptual weighting filter has to be modified in order to suit wideband signals, and pre-emphasis techniques which boost the high frequency regions become important to reduce the dynamic range, yielding a simpler fixed-point implementation, and to ensure a better encoding of the higher frequency contents of the signal. Further, the pitch contents in the spectrum of voiced segments in wideband signals do not extend over the whole spectrum range, and the amount of voicing shows more variation compared to narrow-band signals. Therefore, in case of wideband signals, existing pitch search structures are not very efficient Thus, it is important to improve the closed-loop pitch analysis to better accommodate the variations in the voicing level.

OBJECTS OF THE INVENTION

An object of the present invention is therefore to provide a method and device for efficiently encoding wideband (7000 Hz) sound signals using CELP-type encoding techniques, using improved pitch analysis in order to obtain high a quality reconstructed sound signal.

SUMMARY OF THE INVENTION

More specifically, in accordance with the present invention, there is provided a method for selecting an optimal set of pitch codebook parameters associated to a signal path, from at least two signal paths, having the lowest calculated pitch prediction error. The pitch prediction error is calculated in response to a pitch codevector from a pitch codebook search device. In at least one of the two signal paths, the pitch prediction error is filtered before supplying the pitch codevector for calculation of said pitch prediction error of said one path. Finally, the pitch prediction errors calculated in said at least two signal paths are compared, the signal path having the lowest calculated pitch prediction error is chosen, and the set of pitch codebook parameters associated to the choosen signal path are selected.

The pitch analysts device of the invention, for producing an optimal set of pitch codebook parameters, comprises:

a) at least two signal paths associated to respective sets of pitch codebook parameters, wherein:
   i) each signal path comprises a pitch prediction error calculating device for calculating a pitch prediction error of a pitch codevector from a pitch codebook search device; and
   ii) at least one of the two paths comprises a filter for filtering the pitch codevector before supplying the pitch codevector to the path's pitch prediction error calculating device; and
b) a selector for comparing the pitch prediction errors calculated in the signal paths, for choosing the signal path having the lowest calculated pitch prediction error, and for selecting the set of pitch codebook parameters associated to the choosen signal path.

The new method and device which achieve efficient modeling of the harmonic structure of the speech spectrum uses several forms of low pass filters applied to the past excitation and the one yielding higher prediction gain is selected. When subsample pitch resolution is used, the low pass filters can be incorporated into the interpolation filters used to obtain the higher pitch resolution.

In a preferred embodiment of the invention, each pitch prediction error calculating device of the pitch analysis device described above comprises:

a) a convolution unit for convolving the pitch codevector with a weighted synthesis filter impulse response signal and therefore calculating a convolved pitch codevector;
b) a pitch gain calculator for calculating a pitch gain in response to the convolved pitch codevector and a pitch search target vector;
c) an amplifier for multiplying the convolved pitch codevector by the pitch gain to thereby produce an amplified convolved pitch codevector; and
d) a combiner circuit for combining the amplified convolved pitch codevector with the pitch search target vector to thereby produce the pitch prediction error.

In another preferred embodiment of the invention, the pitch gain calculator comprises a means for calculating said pitch gain $b^{(j)}$ using the relation:

$$b^{(j)}=x^t y^{(j)}/\|y^{(j)}\|^2$$

where j=0, 1, 2, . . . , K, and K corresponds to a number of signal paths,

and where x is said pitch search target vector, and $y^{(j)}$ is said convolved pitch codevector.

The present invention further relates to an encoder, having the pitch analysis device described above, for encoding a wideband input signal and comprising:

a) a linear prediction synthesis filter calculator responsive to the wideband signal for producing linear prediction synthesis filter coefficients;
b) a perceptual weighting filter, responsive to the wideband signal and the linear prediction synthesis filter coefficients, for producing a perceptually weighted signal;
c) an impulse response generator responsive to the linear prediction synthesis filter coefficients for producing a weighted synthesis filter impulse response signal;
d) a pitch search unit for producing pitch codebook parameters, comprising:
   i) a pitch codebook search device responsive to the perceptually weighted signal and the linear prediction synthesis filter coefficients for producing the pitch codevector and an innovative search target vector, and
   ii) the pitch analysis device responsive to the pitch codevector for selecting, from the sets of pitch codebook parameters, the set of pitch codebook parameters associated to the path having the lowest calculated pitch prediction error;
d) an innovative codebook search device, responsive to the weighted synthesis filter impulse response signal, and the innovative search target vector, for producing innovative codebook parameters; and
e) a signal forming device for producing an encoded wideband signal comprising the set of pitch codebook parameters associated to the path having the lowest pitch prediction error, the innovative codebook parameters, and the linear prediction synthesis filter coefficients.

The present invention still further relates to a cellular communication system, a cellular mobile transmitter/receiver unit, a cellular network element, and a bidirectional wireless communication sub-system comprising the above described decoder.

The objects, advantages and other features of the present invention will become more apparent upon reading of the following non restrictive description of a preferred embodiment thereof, given by way of example only with reference to the accompanying drawings.

BRIEF DESCRIPTION OF THE DRAWINGS

In the appended drawings:

FIG. **1** is a schematic block diagram of a preferred embodiment of wideband encoding device;

FIG. **2** is a schematic block diagram of a preferred embodiment of wideband decoding device;

FIG. **3** is a schematic block diagram of a preferred embodiment of pitch analysis device; and

FIG. **4** is a simplified, schematic block diagram of a cellular communication system in which the wideband encoding device of FIG. **1** and the wideband decoding device of FIG. **2** can be used.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENT

As well known to those of ordinary skill in the art, a cellular communication system such as **401** (see FIG. **4**) provides a telecommunication service over a large geographic area by dividing that large geographic area into a number C of smaller cells. The C smaller cells are serviced by respective cellular base stations **402_1**, **402_2** . . . **402_C** to provide each cell with radio signaling, audio and data channels.

Radio signalling channels are used to page mobile radiotelephones (mobile transmitter/receiver units) such as **403** within the limits of the coverage area (cell) of the cellular base station **402**, and to place calls to other radiotelephones **403** located either inside or outside the base station's cell or to another network such as the Public Switched Telephone Network (PSTN) **404**.

Once a radiotelephone **403** has successfully placed or received a call, an audio or data channel is established between this radiotelephone **403** and the cellular base station **402** corresponding to the cell in which the radiotelephone **403** is situated, and communication between the base station **402** and radiotelephone **403** is conducted over that audio or data channel. The radiotelephone **403** may also receive control or timing information over a signalling channel while a call is in progress.

If a radiotelephone **403** leaves a cell and enters another adjacent cell while a call is in progress, the radiotelephone **403** hands over the call to an available audio or data channel of the new cell base station **402**. If a radiotelephone **403** leaves a cell and enters another adjacent cell while no call is in progress, the radiotelephone **403** sends a control message over the signalling channel to log into the base station **402** of the new cell. In this manner mobile communication over a wide geographical area is possible.

The cellular communication system **401** further comprises a control terminal **405** to control communication between the cellular base stations **402** and the PSTN **404**, for example during a communication between a radiotelephone **403** and the PSTN **404**, or between a radiotelephone **403** located in a first cell and a radiotelephone **403** situated in a second cell.

Of course, a bidirectional wireless radio communication subsystem is required to establish an audio or data channel between a base station **402** of one cell and a radiotelephone **403** located in that cell. As illustrated in very simplified form in FIG. **4**, such a bidirectional wireless radio communication subsystem typically comprises in the radiotelephone **403**:

- a transmitter **406** including:
  - an encoder **407** for encoding the voice signal; and
  - a transmission circuit **408** for transmitting the encoded voice signal from the encoder **407** through an antenna such as **409**; and
- a receiver **410** including:
  - a receiving circuit **411** for receiving a transmitted encoded voice signal usually through the same antenna **409**; and
  - a decoder **412** for decoding the received encoded voice signal from the receiving circuit **411**.

The radiotelephone further comprises other conventional radiotelephone circuits **413** to which the encoder **407** and decoder **412** are connected and for processing signals therefrom, which circuits **413** are well known to those of ordinary skill in the art and, accordingly, will not be further described in the present specification.

Also, such a bidirectional wireless radio communication subsystem typically comprises in the base station **402**:

- a transmitter **414** including:
  - an encoder **415** for encoding the voice signal; and
  - a transmission circuit **416** for transmitting the encoded voice signal from the encoder **415** through an antenna such as **417**; and
- a receiver **418** including:
  - a receiving circuit **419** for receiving a transmitted encoded voice signal through the same antenna **417** or through another antenna (not shown); and
  - a decoder **420** for decoding the received encoded voice signal from the receiving circuit **419**.

The base station **402** further comprises, typically, a base station controller **421**, along with its associated database **422**, for controlling communication between the control terminal **405** and the transmitter **414** and receiver **418**.

As well known to those of ordinary skill in the art, voice encoding is required in order to reduce the bandwidth necessary to transmit sound signal, for example voice signal such as speech, across the bidirectional wireless radio communication subsystem, i.e., between a radiotelephone **403** and a base station **402**.

LP voice encoders (such as **415** and **407**) typically operating at 13 kbits/second and below such as Code-Excited Linear Prediction (CELP) encoders typically use a LP synthesis filter to model the short-term spectral envelope of the voice signal. The LP information is transmitted, typically, every 10 or 20 ms to the decoder (such **420** and **412**) and is extracted at the decoder end.

The novel techniques disclosed in the present specification may apply to different LP-based coding systems. However, a CELP-type coding system is used in the preferred embodiment for the purpose of presenting a non-limitative illustration of these techniques. In the same manner, such techniques can be used with sound signals other than voice and speech as well with other types of wideband signals.

FIG. **1** shows a general block diagram of a CELP-type speech encoding device **100** modified to better accommodate wideband signals.

The sampled input speech signal **114** is divided into successive L-sample blocks called "frames". In each frame, different parameters representing the speech signal in the frame are computed, encoded, and transmitted. LP parameters representing the LP synthesis filter are usually computed once every frame. The frame is further divided into smaller blocks of N samples (blocks of length N), in which excitation parameters (pitch and innovation) are determined in the CELP literature, these blocks of length N are called "subframes" and the N-sample signals in the subframes are referred to as N-dimensional vectors. In this preferred embodiment, the length N corresponds to 5 ms while the length L corresponds to 20 ms, which means that a frame contains four subframes (N=80 at the sampling rate of 16 kHz and 64 after down-sampling to 12.8 kHz). Various N-dimensional vectors occur in the encoding procedure. A list of the vectors which appear in FIGS. **1** and **2** as well as a list of transmitted parameters are given herein below:

List of the Main N-Dimensional Vectors

- s Wideband signal input speech vector (after down-sampling, pre-processing, and preemphasis);
- s_w Weighted speech vector;
- s_0 Zero-input response of weighted synthesis filter;
- s_p Down-sampled pre-processed signal; Oversampled synthesized speech signal;
- s' Synthesis signal before deemphasis;
- s_d Deemphasized synthesis signal;
- s_h Synthesis signal after deemphasis and postprocessing;
- x Target vector for pitch search;
- x' Target vector for innovation search;
- h Weighted synthesis filter impulse response;
- v_T Adaptive (pitch) codebook vector at delay T;
- y_T Filtered pitch codebook vector (v_T convolved with h);
- c_k Innovative codevector at index k (k-th entry from the innovation codebook);
- c_f Enhanced scaled innovation codevector;
- u Excitation signal (scaled innovation and pitch codevectors);
- u' Enhanced excitation;
- z Band-pass noise sequence;
- w' White noise sequence; and
- w Scaled noise sequence.

List of Transmitted Parameters

- STP Short term prediction parameters (defining A(z));
- T Pitch lag (or pitch codebook index);
- b Pitch gain (or pitch codebook gain);
- j Index of the low-pass filter used on the pitch codevector;
- k Codevector index (innovation codebook entry); and
- g Innovation codebook gain.

In this preferred embodiment, the STP parameters are transmitted once per frame and the rest of the parameters are transmitted four times per frame (every subframe).

Encoder Side

The sampled speech signal is encoded on a block by block basis by the encoding device **100** of FIG. **1** which is broken down into eleven modules numbered from **101** to **111**.

The input speech is processed into the above mentioned L-sample blocks called frames.

Referring to FIG. **1**, the sampled input speech signal **114** is down-sampled in a down-sampling module **101**. For example, the signal is down-sampled from 16 kHz down to 12.8 kHz, using techniques well known to those of ordinary skill in the art. Down-sampling down to another frequency can of course be envisaged. Down-sampling increases the coding efficiency, since a smaller frequency bandwidth is encoded. This also reduces the algorithmic complexity since the number of samples in a frame is decreased. The use of down-sampling becomes significant when the bit rate is reduced below 16 kbit/s, although down-sampling is not essential above 16 kbit/s.

After down-sampling, the 320-sample frame of 20 ms is reduced to 256-sample frame (down-sampling ratio of 4/5).

The input frame is then supplied to the optional pre-processing block **102**. Pre-processing block **102** may consist of a high-pass filter with a 50 Hz cut-off frequency. High-pass filter **102** removes the unwanted sound components below 50 Hz.

The down-sampled pre-processed signal is denoted by $s_p(n)$, n=0, 1, 2, . . . , L−1, where L is the length of the frame (256 at a sampling frequency of 12.8 kHz). In a preferred embodiment of the preemphasis filter **103**, the signal $s_p(n)$ is preemphasized using a filter having the following transfer function:

$$P(z)1-\mu z^{-1}$$

where μ is a preemphasis factor with a value located between 0 and 1 (a typical value is μ=0.7). A higher-order filter could also be used. It should be pointed out that high-pass filter **102** and preemphasis filter **103** can be interchanged to obtain more efficient fixed-point implementations.

The function of the preemphasis filter **103** is to enhance the high frequency contents of the input signal. It also reduces the dynamic range of the input speech signal, which renders it more suitable for fixed-point implementation. Without preemphasis, LP analysis in fixed-point using single-precision arithmetic is difficult to implement.

Preemphasis also plays an important role in achieving a proper overall perceptual weighting of the quantization error, which contributes to improved sound quality. This will be explained in more detail herein below.

The output of the preemphasis filter **103** is denoted s(n). This signal is used for performing LP analysts in calculator module **104**. LP analysis is a technique well known to those of ordinary skill in the art. In this preferred embodiment, the autocorrelation approach is used. In the autocorrelation approach, the signal s(n) is first windowed using a Hamming window (having usually a length of the order of 30-40 ms). The autocorrelations are computed from the windowed signal, and Levinson-Durbin recursion is used to compute LP filter coefficients, a_i, where i=1, . . . , p, and where p is the LP order, which is typically 16 in wideband coding. The parameters a_i are the coefficients of the transfer function of the LP filter, which is given by the following relation:

$$A(z)=1+\sum_{i=1}^{p} a_i z^{-1}$$

LP analysis is performed in calculator module **104**, which also performs the quantization and interpolation of the LP filter coefficients. The LP filter coefficients are first transformed into another equivalent domain more suitable for quantization and interpolation purposes. The line spectral pair (LSP) and immitance spectral pair (ISP) domains are two domains in which quantization and interpolation can be efficiently performed. The 16 LP filter coefficients, a_i, can be quantized in the order of 30 to 50 bits using split or multi-stage quantization, or a combination thereof. The purpose of the interpolation is to enable updating the LP filter coefficients every subframe while transmitting them once every frame, which improves the encoder performance without increasing the bit rate. Quantization and interpolation of the LP filter coefficients is believed to be otherwise well known to those of ordinary skill in the art and, accordingly, will not be further described in the present specification.

The following paragraphs will describe the rest of the coding operations performed on a subframe basis. In the following description, the filter A(z) denotes the unquantized interpolated LP filter of the subframe, and the filter $\hat{A}(z)$ denotes the quantized interpolated LP filter of the subframe.

Perceptual Weighting:

In analysis-by-synthesis encoders, the optimum pitch and innovation parameters are searched by minimizing the mean squared error between the input speech and synthesized speech in a perceptually weighted domain. This is equivalent to minimizing the error between the weighted input speech and weighted synthesis speech.

The weighted signal $s_w(n)$ is computed in a perceptual weighting filter **105**. Traditionally, the weighted signal $s_w(n)$ is computed by a weighting filter having a transfer function W(z) in the form:

$$W(z)=A(z/\gamma_1)/A(z/\gamma_2)$$

where

$$0<\gamma_2<\gamma_1\leqq 1$$

As well known to those of ordinary skill in the art, in prior art analysis-by-synthesis (AbS) encoders, analysis shows that the quantization error is weighted by a transfer function $W^{-1}(z)$, which is the inverse of the transfer function of the perceptual weighting filter **105**. This result is well described by B. S. Atal and M. R. Schroeder in "Predictive coding of speech and subjective error criteria", IEEE Transaction ASSP, vol. 27, no. 3, pp. 247-254; June 1979. Transfer function $W^{-1}(z)$ exhibits some of the formant structure of the input speech signal. Thus, the masking property of the human ear is exploited by shaping the quantization error so that it has more energy in the formant regions where it will be masked by the strong signal energy present in these regions. The amount of weighting is controlled by the factors γ_1 and γ_2.

The above traditional perceptual weighting filter **105** works well with telephone band signals. However, it was found that this traditional perceptual weighting filter **105** is not suitable for efficient perceptual weighting of wideband signals. It was also found that the traditional perceptual weighting filter **105** has inherent limitations in modelling the formant structure and the required spectral tilt concurrently. The spectral tilt is more pronounced in wideband signals due to the wide dynamic range between low and high frequencies.

The prior art has suggested to add a tilt filter into W(z) in order to control the tilt and formant weighting of the wideband input signal separately.

A novel solution to this problem is, in accordance with the present invention, to introduce the preemphasis filter **103** at the input, compute the LP filter A(z) based on the preemphasized speech s(n), and use a modified filter W(z) by fixing its denominator.

LP analysis is performed in module **104** on the preemphasized signal s(n) to obtain the LP filter A(z). Also, a new perceptual weighting filter **105** with fixed denominator is used. An example of transfer function for the perceptual weighting filter **104** is given by the following relation:

$$W(z)=A(z/\gamma_1)/(1-\gamma_2 z^{-1})$$

where

$$0<\gamma_2<\gamma_1\leqq 1$$

A higher order can be used at the denominator. This structure substantially decouples the formant weighting from the tilt.

Note that because A(z) is computed based on the preemphasized speech signal s(n), the tilt of the filter $1/A(z/\gamma_1)$ is less pronounced compared to the case when A(z) is computed based on the original speech. Since deemphasis is performed at the decoder end using a filter having the transfer function:

$$P^{-1}(z)=1/(1-\mu z^{-1}),$$

the quantization error spectrum is shaped by a filter having a transfer function $W^{-1}(z)P^{-1}(z)$. When γ_2 is set equal to μ, which is typically the case, the spectrum of the quantization error is shaped by a filter whose transfer function is $1/A(z/\gamma_1)$, with A(z) computed based on the preemphasized speech signal. Subjective listening showed that this structure for achieving the error shaping by a combination of preemphasis and modified weighting filtering is very efficient for encoding wideband signals, in addition to the advantages of ease of fixed-point algorithmic implementation.

Pitch Analysis:

In order to simplify the pitch analysis, an open-loop pitch lag T_{OL} is first estimated in the open-loop pitch search module **106** using the weighted speech signal $s_w(n)$. Then the closed-loop pitch analysis, which is performed in closed-loop pitch search module **107** on a subframe basis, is restricted around the open-loop pitch lag T_{OL} which significantly reduces the search complexity of the LTP parameters T and b (pitch lag and pitch gain). Open-loop pitch analysis is usually performed in module **106** once every 10 ms (two subframes) using techniques well known to those of ordinary skill in the art.

The target vector x for LTP (Long Term Prediction) analysis is first computed. This is usually done by subtracting the zero-input response s_0 of weighted synthesis filter $W(z)/\hat{A}(z)$ from the weighted speech signal $s_w(n)$. This zero-input response s_0 is calculated by a zero-input response calculator **108**. More specifically, the target vector x is calculated using the following relation:

$$x=s_w-s_0$$

where x is the N-dimensional target vector, s_w is the weighted speech vector in the subframe, and s_0 is the zero-input response of filter $W(z)/\hat{A}(z)$ which is the output of the combined filter $W(z)/\hat{A}(z)$ due to its initial states. The zero-input response calculator **108** is responsive to the quantized interpolated LP filter $\hat{A}(z)$ from the LP analysis, quantization and interpolation calculator **104** and to the initial states of the weighted synthesis filter $W(z)/\hat{A}(z)$ stored in memory module **111** to calculate the zero-input response s_0 (that part of the response due to the initial states as determined by setting the inputs equal to zero) of filter $W(z)/\hat{A}(z)$. This operation is well known to those of ordinary skill in the art and, accordingly, will not be further described.

Of course, alternative but mathematically equivalent approaches can be used to compute the target vector x.

A N-dimensional impulse response vector h of the weighted synthesis filter $W(z)/\hat{A}(z)$ is computed in the impulse response generator **109** using the LP filter coefficients A(z) and $\hat{A}(z)$ from module **104**. Again, this operation is well known to those of ordinary skill in the art and, accordingly, will not be further described in the present specification.

The closed-loop pitch (or pitch codebook) parameters b, T and j are computed in the closed-loop pitch search module **107**, which uses the target vector x, the impulse response vector h and the open-loop pitch lag T_{OL} as inputs. Traditionally, the pitch prediction has been represented by a pitch filter having the following transfer function:

$$1/(1-bz^{-T})$$

where b is the pitch gain and T is the pitch delay or lag. In this case, the pitch contribution to the excitation signal u(n) is given by bu(n−T), where the total excitation is given by

$$u(n)=bu(n-T)+gc_k(n)$$

with g being the innovative codebook gain and $c_k(n)$ the innovative codevector at index k.

This representation has limitations if the pitch lag T is shorter than the subframe length N. In another representation, the pitch contribution can be seen as a pitch codebook containing the past excitation signal. Generally, each vector in the pitch codebook is a shift-by-one version of the previous vector (discarding one sample and adding a new sample). For pitch lags T>N, the pitch codebook is equivalent to the filter structure $(1/(1-bz^{-T})$, and a pitch codebook vector $v_T(n)$ at pitch lag T is given by

$$v_T(n)=u(n-T),\ n=0,\ \ldots,\ N-1.$$

For pitch lags T shorter than N, a vector $v_T(n)$ is built by repeating the available samples from the past excitation until the vector is completed (this is not equivalent to the filter structure).

In recent encoders, a higher pitch resolution is used which significantly improves the quality of voiced sound segments. This is achieved by oversampling the past excitation signal using polyphase interpolation filters. In this case, the vector $v_T(n)$ usually corresponds to an interpolated version of the past excitation, with pitch lag T being a non-integer delay (e.g. 50.25).

The pitch search consists of finding the best pitch lag T and gain b that minimize the mean squared weighted error E between the target vector x and the scaled filtered past excitation. Error E being expressed as:

$$E=\|x-by_T\|^2$$

where y_T is the filtered pitch codebook vector at pitch lag T:

$$y_T(n)=v_T(n)*h(n)=\sum_{i=o}^{n} v_T(i)h(n-i),\ n=0,\ \ldots\ ,N-1.$$

It can be shown that the error E is minimized by maximizing the search criterion

$$C = \frac{x^t y_T}{\sqrt{y_T^t y_T}}$$

where t denotes vector transpose.

In the preferred embodiment of the present invention, a ⅓ subsample pitch resolution is used, and the pitch (pitch codebook) search is composed of three stages.

In the first stage, an open-loop pitch lag T_{OL} is estimated in open-loop pitch search module **106** in response to the weighted speech signal $s_w(n)$. As indicated in the foregoing description, this open-loop pitch analysis is usually performed once every 10 ms (two subframes) using techniques well known to those of ordinary skill in the art.

In the second stage, the search criterion C is searched in the closed-loop pitch search module **107** for integer pitch lags around the estimated open-loop pitch lag T_{OL} (usually ±5), which significantly simplifies the search procedure. A simple procedure is used for updating the filtered codevector y_T without the need to compute the convolution for every pitch lag.

Once an optimum integer pitch lag is found in the second stage, a third stage of the search (module **107**) tests the fractions around that optimum integer pitch lag.

When the pitch predictor is represented by a filter of the form $1/(1-bz^{-T})$, which is a valid assumption for pitch lags T>N, the spectrum of the pitch filter exhibits a harmonic structure over the entire frequency range, with a harmonic frequency related to 1/T. In case of wideband signals, this structure is not very efficient since the harmonic structure in wideband signals does not cover the entire extended spectrum. The harmonic structure exists only up to a certain frequency, depending on the speech segment. Thus, in order to achieve efficient representation of the pitch contribution in voiced segments of wideband speech, the pitch prediction filter needs to have the flexibility of varying the amount of periodicity over the wideband spectrum.

A new method which achieves efficient modeling of the harmonic structure of the speech spectrum of wideband signals is disclosed in the present specification, whereby several forms of low pass filters are applied to the past excitation and the low pass filter with higher prediction gain is selected.

When subsample pitch resolution is used, the low pass filters can be incorporated into the interpolation filters used to obtain the higher pitch resolution. In this case, the third stage of the pitch search, in which the fractions around the chosen integer pitch lag are tested, is repeated for the several interpolation filters having different low-pass characteristics and the fraction and filter index which maximize the search criterion C are selected.

A simpler approach is to complete the search in the three stages described above to determine the optimum fractional pitch lag using only one interpolation filter with a certain frequency response, and select the optimum low-pass filter shape at the end by applying the different predetermined low-pass filters to the chosen pitch codebook vector v_T and select the low-pass filter which minimizes the pitch prediction error. This approach is discussed in detail below.

FIG. **3** illustrates a schematic block diagram of a preferred embodiment of the proposed approach.

In memory module **303**, the past excitation signal u(n), n<0, is stored. The pitch codebook search module **301** is responsive to the target vector x, to the open-loop pitch lag T_{OL} and to the past excitation signal u(n), n<0, from memory module **303** to conduct a pitch codebook (pitch codebook) search minimizing the above-defined search criterion C. From the result of the search conducted in module **301**, module **302** generates the optimum pitch codebook vector v_T. Note that since a sub-sample pitch resolution is used (fractional pitch), the past excitation signal u(n), n<0, is interpolated and the pitch codebook vector v_T corresponds to the interpolated past excitation signal. In this preferred embodiment, the interpolation filter (in module **301**, but not shown) has a low-pass filter characteristic removing the frequency contents above 7000 Hz.

In a preferred embodiment, K filter characteristics are used; these filter characteristics could be low-pass or band-pass filter characteristics. Once the optimum codevector v_T is determined and supplied by the pitch codevector generator **302**, K filtered versions of v_T are computed respectively using K different frequency shaping filters such as $\mathbf{305}^{(j)}$, where j=1, 2, . . . , K. These filtered versions are denoted $v_f^{(j)}$, where j=1, 2, . . . , K. The different vectors $v_f^{(j)}$ are convolved in respective modules $\mathbf{304}^{(j)}$, where j=0, 1, 2, . . . K, with the impulse response h to obtain the vectors $y^{(j)}$, where j=0, 1, 2, . . . , K. To calculate the mean squared pitch prediction error for each vector $y^{(j)}$, the value $y^{(j)}$ is multiplied by the gain b by means of a corresponding amplifier $\mathbf{307}^{(j)}$ and the value $by^{(j)}$ is subtracted from the target vector x by means of a corresponding subtracter $\mathbf{308}^{(j)}$. Selector **309** selects the frequency shaping filter $\mathbf{305}^{(j)}$ which minimizes the mean squared pitch prediction error

$$e^{(j)}=\|x-b^{(j)}y^{(j)}\|^2$$

$$j=1, 2, \ldots, K$$

To calculate the mean squared pitch prediction error $e^{(j)}$ for each value of $y^{(j)}$ the value $y^{(j)}$ is multiplied by the gain b by means of a corresponding amplifier $\mathbf{307}^{(j)}$ and the value $b^{(j)}y^{(j)}$ is subtracted from the target vector x by means of subtracters $\mathbf{308}^{(j)}$. Each gain $b^{(j)}$ is calculated in a corresponding gain calculator $\mathbf{306}^{(j)}$ in association with the frequency shaping filter at index j, using the following relationship:

$$b^{(j)}=x^t y^{(j)}/\|y^{(j)}\|^2$$

In selector **309**, the parameters b, T, and j are chosen based on v_T or $v_f^{(j)}$ which minimizes the mean squared pitch prediction error e.

Referring back to FIG. **1**, the pitch codebook index T is encoded and transmitted to multiplexer **112**. The pitch gain b is quantized and transmitted to multiplexer **112**. With this new approach, extra information is needed to encode the index j of the selected frequency shaping filter in multiplexer **112**. For example, if three filters are used (j=0, 1, 2, 3), then two bits are needed to represent this information. The filter index information j can also be encoded jointly with the pitch gain b.

Innovative Codebook Search:

Once the pitch, or LTP (Long Term Prediction) parameters b, T, and j are determined, the next step is to search for the optimum innovative excitation by means of search module **110** of FIG. **1**. First, the target vector x is updated by subtracting the LTP contribution:

$$x'=x-by_T$$

where b is the pitch gain and y_T is the filtered pitch codebook vector (the past excitation at delay T filtered with the selected low pass filter and convolved with the inpulse response h as described with reference to FIG. **3**).

The search procedure in CELP is performed by finding the optimum excitation codevector c_k and gain g which minimize the mean-squared error between the target vector and the scaled filtered codevector

$$E=\|x'-gHc_k\|^2$$

where H is a lower triangular convolution matrix derived from the impulse response vector h.

In the preferred embodiment of the present invention, the innovative codebook search is performed in module **110** by means of an algebraic codebook as described in U.S. Pat. No. 5,444,816 (Adoul et al.) issued on Aug. 22, 1995; U.S. Pat. No. 5,699,482 granted to Adoul et al., on Dec. 17, 1997; U.S. Pat. No. 5,754,976 granted to Adoul et al., on May 19, 1998; and U.S. Pat. No. 5,701,392 (Adoul et al.) dated Dec. 23, 1997.

Once the optimum excitation codevector c_k and its gain g are chosen by module **110**, the codebook index k and gain g are encoded and transmitted to multiplexer **112**.

Referring to FIG. **1**, the parameters b, T, j, $\hat{A}(z)$, k and g are multiplexed through the multiplexer **112** before being transmitted through a communication channel.

Memory Update:

In memory module **111** (FIG. **1**), the states of the weighted synthesis filter $W(z)/\hat{A}(z)$ are updated by filtering the excitation signal $u=gc_k+bv_T$ through the weighted synthesis filter. After this filtering, the states of the filter are memorized and used in the next subframe as initial states for computing the zero-input response in calculator module **108**.

As in the case of the target vector x, other alternative but mathematically equivalent approaches well known to those of ordinary skill in the art can be used to update the filter states.

Decoder Side

The speech decoding device **200** of FIG. **2** illustrates the various steps carried out between the digital input **222** (input stream to the demultiplexer **217**) and the output sampled speech **223** (output of the adder **221**).

Demultiplexer **217** extracts the synthesis model parameters from the binary information received from a digital input channel. From each received binary frame, the extracted parameters are:

- the short-term prediction parameters (STP) $\hat{A}(z)$ (once per frame);
- the long-term prediction (LTP) parameters T, b, and j (for each subframe); and
- the innovation codebook index k and gain g (for each subframe).

The current speech signal is synthesized based on these parameters as will be explained hereinbelow.

The innovative codebook **218** is responsive to the index k to produce the innovation codevector c_k, which is scaled by the decoded gain factor g through an amplifier **224**. In the preferred embodiment, an innovative codebook **218** as described in the above mentioned U.S. Pat. Nos. 5,444,816; 5,699,482; 5,754,976; and 5,701,392 is used to represent the innovative codevector c_k.

The generated scaled codevector gc_k at the output of the amplifier **224** is processed through a innovation filter **205**.

Periodicity Enhancement:

The generated scaled codevector at the output of the amplifier **224** is processed through a frequency-dependent pitch enhancer **205**.

Enhancing the periodicity of the excitation signal u improves the quality in case of voiced segments. This was done in the past by filtering the innovation vector from the innovative codebook (fixed codebook) **218** through a filter in the form $1/(1-\epsilon bz^{-T})$ where ϵ is a factor below 0.5 which controls the amount of introduced periodicity. This approach is less efficient in case of wideband signals since it introduces periodicity over the entire spectrum. A new alternative approach, which is part of the present invention, is disclosed whereby periodicity enhancement is achieved by filtering the innovative codevector c_k from the innovative (fixed) codebook through an innovation filter **205** (F(z)) whose frequency response emphasizes the higher frequencies more than lower frequencies. The coefficients of F(z) are related to the amount of periodicity in the excitation signal u.

Many methods known to those skilled in the art are available for obtaining valid periodicity coefficients. For example, the value of gain b provides an indication of periodicity. That is, if gain b is close to 1, the periodicity of the excitation signal u is high, and if gain b is less than 0.5, then periodicity is low.

Another efficient way to derive the filter F(z) coefficients used in a preferred embodiment, is to relate them to the amount of pitch contribution in the total excitation signal u. This results in a frequency response depending on the subframe periodicity, where higher frequencies are more strongly emphasized (stronger overall slope) for higher pitch gains. Innovation filter **205** has the effect of lowering the energy of the innovative codevector c_k at low frequencies when the excitation signal u is more periodic, which enhances the periodicity of the excitation signal u at lower frequencies more than higher frequencies. Suggested forms for innovation filter **205** are

$$F(z)=1-\sigma z^{-1}, \qquad (1)$$

or

$$F(z)=-\alpha z+1-\alpha z^{-1} \qquad (2)$$

where σ or α are periodicity factors derived from the level of periodicity of the excitation signal u.

The second three-term form of F(z) is used in a preferred embodiment. The periodicity factor α is computed in the voicing factor generator **204**. Several methods can be used to derive the periodicity factor α based on the periodicity of the excitation signal u. Two methods are presented below.

Method 1:

The ratio of pitch contribution to the total excitation signal u is first computed in voicing factor generator **204** by

$$R_p=\frac{b^2 v_T^t v_T}{u^t u}=\frac{b^2\sum_{n=0}^{N-1} v_T^2(n)}{\sum_{n=0}^{N-1} u^2(n)}$$

where v_T is the pitch codebook vector, b is the pitch gain, and u is the excitation signal u given at the output of the adder **219** by

$$u=gc_k+bv_T$$

Note that the term bv_T has its source in the pitch codebook (pitch codebook) **201** in response to the pitch lag T and the past value of u stored in memory **203**. The pitch codevector v_T from the pitch codebook **201** is then processed through a low-pass filter **202** whose cut-off frequency is adjusted by means of the index j from the demultiplexer **217**. The resulting codevector y_T is then multiplied by the gain b from the demultiplexer **217** through an amplifier **226** to obtain the signal bv_T.

The factor α is calculated in voicing factor generator **204** by $\alpha=qR_p$ bounded by $\alpha<q$ where q is a factor which controls the amount of enhancement (q is set to 0.25 rn this preferred embodiment).

Method 2:

Another method used in a preferred embodiment of the invention for calculating periodicity factor α is discussed below.

First, a voicing factor r_v is computed in voicing factor generator **204** by

$$r_v=(E_v-E_c)/(E_v+E_c)$$

where E_v is the energy of the scaled pitch codevector bv_T and E_c is the energy of the scaled innovative codevector gc_k. That is

$$E_v = b^2 v_T^t v_T = b^2 \sum_{n=0}^{N-1} v_T^2(n)$$

and

$$E_c = g^2 c_k^t c_k = g^2 \sum_{n=0}^{N-1} c_k^2(n)$$

Note that the value of r_v lies between −1 and 1 (1 corresponds to purely voiced signals and −1 corresponds to purely unvoiced signals).

In this preferred embodiment, the factor α is then computed in voicing factor generator **204** by

$$\alpha=0.125(1+r_v)$$

which corresponds to a value of 0 for purely unvoiced signals and 0.25 for purely voiced signals.

In the first, two-term form of F(z), the periodicity factor σ can be approximated by using $\sigma=2\alpha$ in methods 1 and 2 above. In such a case, the periodicity factor σ is calculated as follows in method 1 above:

$$\sigma=2qR_p \text{ bounded by } \sigma<2q.$$

In method 2, the periodicity factor σ is calculated as follows:

$$\sigma=0.25(1+r_v).$$

The enhanced signal c_f is therefore computed by filtering the scaled innovative codevector gc_k through the innovation filter **205** (F(z)).

The enhanced excitation signal u' is computed by the adder **220** as:

$$u'=c_f+bv_T$$

Note that this process is not performed at the encoder **100**. Thus, it is essential to update the content of the pitch codebook **201** using the excitation signal u without enhancement to keep synchronism between the encoder **100** and decoder **200**. Therefore, the excitation signal u is used to update the memory **203** of the pitch codebook **201** and the enhanced excitation signal u' is used at the input of the LP synthesis filter **206**.

Synthesis and Deemphasis

The synthesized signal s' is computed by filtering the enhanced excitation signal u' through the LP synthesis filter **206** which has the form $1/\hat{A}(z)$, where $\hat{A}(z)$ is the interpolated LP filter in the current subframe. As can be seen in FIG. **2**, the quantized LP coefficients $\hat{A}(z)$ online **225** from demultiplexer **217** are supplied to the LP synthesis filter **206** to adjust the parameters of the LP synthesis filter **206** accordingly. The deemphasis filter **207** is the inverse of the preemphasis filter **103** of FIG. **1**. The transfer function of the deemphasis filter **207** is given by

$$D(z)=1/(1-\mu z^{-1})$$

where μ is a preemphasis factor with a value located between 0 and 1 (a typical value is $\mu=0.7$). A higher-order filter could also be used.

The vector s' is filtered through the deemphasis filter D(z) (module **207**) to obtain the vector s_d which is passed through the high-pass filter **208** to remove the unwanted frequencies below 50 Hz and further obtain s_h.

Oversampling and High-Frequency Regeneration

The over-sampling module **209** conducts the inverse process of the down-sampling module **101** of FIG. **1**. In this preferred embodiment, oversampling converts from the 12.8 kHz sampling rate to the original 16 kHz sampling rate, using techniques well known to those of ordinary skill in the art. The oversampled synthesis signal is denoted $\hat{S}$. Signal $\hat{S}$ is also referred to as the synthesized wideband intermediate signal.

The oversampled synthesis $\hat{S}$ signal does not contain the higher frequency components which were lost by the down-sampling process (module **101** of FIG. **1**) at the encoder **100**. This gives a low-pass perception to the synthesized speech signal. To restore the full band of the original signal, a high frequency generation procedure is disclosed. This procedure is performed in modules **210** to **216**, and adder **221**, and requires input from voicing factor generator **204** (FIG. **2**).

In this new approach, the high frequency contents are generated by filling the upper part of the spectrum with a white noise properly scaled in the excitation domain, then converted to the speech domain, preferably by shaping it with the same LP synthesis filter used for synthesizing the down-sampled signal $\hat{S}$.

The high frequency generation procedure in accordance with the present invention is described hereinbelow.

The random noise generator **213** generates a white noise sequence w' with a flat spectrum over the entire frequency bandwidth, using techniques well known to those of ordinary skill in the art. The generated sequence is of length N' which is the subframe length in the original domain. Note that N is the subframe length in the down-sampled domain. In this preferred embodiment, N=64 and N'=80 which correspond to 5 ms.

The white noise sequence is properly scaled in the-gain adjusting module **214**. Gain adjustment comprises the following steps. First, the energy of the generated noise sequence w' is set equal to the energy of the enhanced excitation signal u' computed by an energy computing module **210**, and the resulting scaled noise sequence is given by

$$w(n) = x'(n)\sqrt{\frac{\sum_{n=0}^{N-1} u'^2(n)}{\sum_{n=0}^{N'-1} w'^2(n)}},\ n = 0, \ldots , N'-1.$$

The second step in the gain scaling is to take into account the high frequency contents of the synthesized signal at the output of the voicing factor generator **204** so as to reduce the energy of the generated noise in case of voiced segments (where less energy is present at high frequencies compared to unvoiced segments). In this preferred embodiment, measuring the high frequency contents is implemented by measuring the tilt of the synthesis signal through a spectral tilt calculator **212** and reducing the energy accordingly. Other measurements such as zero crossing measurements can equally be used. When the tilt is very strong, which corresponds to voiced segments, the noise energy is further reduced. The tilt factor is computed in module **212** as the first correlation coefficient of the synthesis signal s_h and it is given by:

$$\text{tilt} = \frac{\sum_{n=1}^{N-1} s_h(n)s_h(n-1)}{\sum_{n=0}^{N-1} s_h^2(n)},$$

conditioned by tilt $\geqq 0$ and tilt $\geqq r_v$.
where voicing factor r_v is given by

$$r_v=(E_v-E_c)/(E_v+E_c)$$

where E_v is the energy of the scaled pitch codevector bv_T and E_c is the energy of the scaled innovative codevector gc_k as described earlier. Voicing factor r_v is most often less than tilt but this condition was introduced as a precaution against high frequency tones where the tilt value is negative and the value of r_v is high. Therefore, this condition reduces the noise energy for such tonal signals.

The tilt value is 0 in case of flat spectrum and 1 in case of strongly voiced signals, and it is negative in case of unvoiced signals where more energy is present at high frequencies.

Different methods can be used to derive the scaling factor g_t from the amount of high frequency contents. In this invention, two methods are given based on the tilt of signal described above.

Method 1:

The scaling factor g_t is derived from the tilt by

$$g_t=1-\text{tilt bounded by } 0.2\leqq g_t\leqq 1.0$$

For strongly voiced signal where the tilt approaches 1, g_t is 0.2 and for strongly unvoiced signals g_t becomes 1.0.

Method 2:

The tilt factor g_t is first restricted to be larger or equal to zero, then the scaling factor is derived from the tilt by

$$g_t=10^{-0.6\,tilt}$$

The scaled noise sequence w_g produced in gain adjusting module **214** is therefore given by:

$$w_g=g_t w.$$

When the tilt is close to zero, the scaling factor g_t is close to 1, which does not result in energy reduction. When the tilt value is 1, the scaling factor g_t results in a reduction of 12 dB in the energy of the generated noise.

Once the noise is properly scaled (w_g), it is brought into the speech domain using the spectral shaper **215**. In the preferred embodiment, this is achieved by filtering the noise w_g through a bandwidth expanded version of the same LP synthesis filter used in the down-sampled domain ($1/\hat{A}(z/0.8)$). The corresponding bandwidth expanded LP filter coefficients are calculated in spectral shaper **215**.

The filtered scaled noise sequence w_f is then band-pass filtered to the required frequency range to be restored using the band-pass filter **216**. In the preferred embodiment, the band-pass filter **216** restricts the noise sequence to the frequency range 5.6-7.2 kHz. The resulting band-pass filtered noise sequence z is added in adder **221** to the oversampled synthesized speech signal $\hat{S}$ to obtain the final reconstructed sound signal s_{out} on the output **223**.

Although the present invention has been described hereinabove by way of a preferred embodiment thereof, this embodiment can be modified at will, within the scope of the appended claims, without departing from the spirit and nature of the subject invention. Even though the preferred embodiment discusses the use of wideband speech signals, it will be obvious to those skilled in the art that the subject invention is also directed to other embodiments using wideband signals in general and that it is not necessarily limited to speech applications.

What is claimed is:

1. A method comprising:

receiving a pitch codevector and a target vector;

passing the pitch codevector through at least one frequency shaping filter to obtain at least one filtered version of the pitch codevector;

determining at least two pitch prediction errors, a prediction error being representative of a difference between the target vector and a product of the pitch codevector and a pitch gain value, at least one of the pitch prediction errors being determined from a filtered version of the pitch codevector;

comparing, via a processor, the at least two pitch prediction errors to identify the pitch prediction error having a lowest error energy value;

selecting the pitch gain associated with the lowest energy value; and

providing an indication of the selected pitch gain.

2. The method according to claim **1**, further comprising providing an indication of the at least one frequency shaping filter used to obtain the at least one filtered version of the pitch codevector.

3. The method according to claim **2**, wherein the indication of the at least one frequency shaping filter is an index representative of the at least one frequency shaping filter.

4. The method according to claim **1**, wherein passing the pitch codevector through the at least one frequency shaping filter comprises passing the pitch codevector through more than one frequency shaping filter to obtain more than one filtered version of the pitch codevector.

5. The method according to claim **1**, wherein passing the pitch code vector through the at least one frequency shaping filter comprises passing the pitch code vector through a low-pass filter.

6. The method according to claim **1**, wherein passing the pitch code vector through the at least one frequency shaping filter comprises passing the pitch code vector through a band-pass filter.

7. The method according to claim **1**, wherein determining the at least two pitch errors comprises:

convolving the pitch codevector with a weighted synthesis filter impulse response signal to obtain a convolved pitch codevector;

multiplying the convolved pitch codevector by the pitch gain value to produce an amplified convolved pitch codevector; and

subtracting the amplified convolved pitch codevector from the target vector.

8. The method according to claim **7**, further comprising calculating the pitch gain value $b^{(j)}$ from the convolved pitch codevector using the relation

$$b^{(j)}=x^t y^{(j)}/\|y^{(j)}\|^2$$

where j=0, 1, 2 . . . , K, and K corresponds to a number of pitch prediction errors to be determined, x is the target vector and $y^{(j)}$ is the convolved pitch codevector.

9. The method according to claim **1**, wherein the pitch codevector is an interpolated pitch codevector having sub-sample pitch resolution.

10. A device comprising:

at least one frequency shaping filter for obtaining at least one filtered version of a pitch codevector;

at least one subtractor for determining at least two pitch prediction errors, a prediction error being representative of a difference between a target vector and a product of the pitch codevector and a pitch gain value, at least one of the pitch prediction errors being determined from a filtered version of the pitch codevector;

a processor for comparing the at least two pitch prediction errors to identify the pitch prediction error having a lowest error energy value; and

a selector for selecting the pitch gain associated with the lowest energy value and for providing an indication of the selected pitch gain.

11. The device according to claim **10**, wherein the selector further provides an indication of the at least one frequency shaping filter used to obtain the at least one filtered version of the pitch codevector.

12. The device according to claim **11**, wherein the indication of the at least one frequency shaping filter is an index representative of the at least one frequency shaping filter.

13. The device according to claim **10**, comprising more than one frequency shaping filter to obtain more than one filtered version of the pitch codevector.

14. The device according to claim **10**, wherein the at least one frequency shaping filter comprises a low-pass filter for filtering the pitch code vector.

15. The device according to claim **10**, wherein the at least one frequency shaping filter comprises a band-pass filter for filtering the pitch code vector.

16. The device according to claim **10**, comprising:

a convolution module for convolving the pitch codevector with a weighted synthesis filter impulse response signal to obtain a convolved pitch codevector; and

an amplifier for multiplying the convolved pitch codevector by the pitch gain value to produce an amplified convolved pitch codevector;

wherein the at least one subtractor further subtracts the amplified convolved pitch codevector from the target vector.

17. The device according to claim **16**, further comprising a gain calculator for calculating the pitch gain value $b^{(j)}$ from the convolved pitch codevector using the relation

$$b^{(j)}=x^t y^{(j)}/\|y^{(j)}\|^2$$

where j=0, 1, 2 . . . , K, and K corresponds to a number of pitch prediction errors to be determined, x is the target vector and $y^{(j)}$ is the convolved pitch codevector.

18. The device according to claim **10**, wherein the pitch codevector is an interpolated pitch codevector having sub-sample pitch resolution.

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