

BTSC ENCODER**Field of the Invention**

The present invention relates generally to stereophonic audio encoders used for television broadcasting. More particularly, the invention relates to a digital encoder for generating the audio signals used in the broadcast of stereophonic television signals in the United States and in other countries.

Background of the Invention

In the 1980's, the United States Federal Communications Commission (FCC) adopted new regulations covering the audio portion of television signals which permitted television programs to be broadcast and received with bichannel audio, e.g., stereophonic sound. In those regulations, the FCC recognized and gave special protection to a method of broadcasting additional audio channels endorsed by the Electronic Industries Association and the National Association of Broadcasters and called the Broadcast Television Systems Committee (BTSC) system. This well-known standard is sometimes referred to as Multichannel Television Sound (MTS) and is described in the FCC document entitled, MULTICHANNEL TELEVISION SOUND TRANSMISSION AND AUDIO PROCESSING REQUIREMENTS FOR THE BTSC SYSTEM (OET Bulletin No. 60, Revision A, February 1986), as well as in the document published by the Electronic Industries Association entitled, MULTICHANNEL TELEVISION SOUND BTSC SYSTEM RECOMMENDED PRACTICES (EIA Television Systems Bulletin No. 5, July 1985). Television signals generated according to the BTSC standard are referred to hereinafter as "BTSC signals".

The original monophonic television signals carried only a single channel of audio. Due to the configuration of the monophonic television signal and the need to maintain compatibility with existing television sets, the stereophonic information was necessarily located in a higher frequency region of the BTSC signal making the stereophonic channel much noisier than the monophonic audio channel. This resulted in an inherently higher noise floor for the stereo signal than for the monophonic signal. The BTSC standard overcame this problem by

1 defining an encoding system that provided additional signal processing for the
2 stereophonic audio signal. Prior to broadcast of a BTSC signal by a television
3 station, the audio portion of a television program is encoded in the manner
4 prescribed by the BTSC standard, and upon reception of a BTSC signal a receiver
5 (e.g., a television set) then decodes the audio portion in a complementary
6 manner. This complementary encoding and decoding insures that the signal-to-
7 noise ratio of the entire stereo audio signal is maintained at acceptable levels.

8 Figure 1 is a block diagram of a prior art BTSC encoding system, or more
9 simply, a BTSC encoder 100, as defined by the BTSC standard. Encoder 100
10 receives left and right channel audio input signals (indicated in Figure 1 as "L"
11 and "R", respectively) and generates therefrom a conditioned sum signal and an
12 encoded difference signal. It should be appreciated that while the system of the
13 prior art and that of the present invention is described as useful for encoding the
14 left and right audio signals of a stereophonic signal that is subsequently
15 transmitted as a television signal, the BTSC system also provides means to encode
16 a separate audio signal, e.g., audio information in a different language, which is
17 separated and selected by the end receiver. Further, noise reduction components
18 of the BTSC encoding system can be used for other purposes besides television
19 broadcast, such as for improving audio recordings.

20 System 100 includes an input section 110, a sum channel processing
21 section 120, and a difference channel processing section 130. Input section 110
22 receives the left and right channel audio input signals and generates therefrom a
23 sum signal (indicated in Figure 1 as "L+R") and a difference signal (indicated in
24 Figure 1 as "L-R"). It is well known that for stereophonic signals, the sum
25 signal L+R may be used by itself to provide monophonic audio reproduction and
26 it is this signal that is decoded by existing monophonic audio television sets to
27 reproduce sound. In stereophonic sets, the sum and difference signals can be
28 added to and subtracted from one another to recover the original two stereophonic
29 signals (L) and (R). Input section 110 includes two signal adders 112, 114.
30 Adder 112 sums the left and right channel audio input signals to generate the sum
31 signal, and adder 114 subtracts the right channel audio input signal from the left
32 channel audio input signal to generate the difference signal. As described above,

1 the sum signal $L+R$ is transmitted through a transmission media with the same
2 signal to noise ratio as achieved with the prior monophonic signals. However,
3 the difference signal $L-R$ is transmitted through a very noisy channel, particularly
4 at the higher frequency portion of the relevant spectrum so that the decoded
5 difference signal has a poorer signal-to-noise ratio because of the noisy medium
6 and reduced dynamic range of the medium. The dynamic range is defined as the
7 range of signals between the level of the noise floor and the maximum level
8 where signal saturation occurs. In the difference signal channel the dynamic
9 range decreases at higher frequencies. Accordingly, the difference signal is
10 subjected to additional processing than that of the sum signal so that the dynamic
11 range can be substantially preserved.

12 More particularly, the sum channel processing section 120 receives the
13 sum signal and generates therefrom the conditioned sum signal. Section 120
14 includes a $75\mu\text{s}$ preemphasis filter 122 and a bandlimiter 124. The sum signal is
15 applied to the input of filter 122 which generates therefrom an output signal that
16 is applied to the input of bandlimiter 124. The output signal generated by the
17 latter is then the conditioned sum signal.

18 The difference channel processing section 130 receives the difference
19 signal and generates therefrom the encoded difference signal. Section 130
20 includes a fixed preemphasis filter 132 (shown implemented as a cascade of two
21 filters 132a and 132b), a variable gain amplifier 134 preferably in the form of a
22 voltage-controlled amplifier, a variable preemphasis/deemphasis filter (referred to
23 hereinafter as a "variable emphasis filter") 136, an overmodulation protector and
24 bandlimiter 138, a fixed gain amplifier 140, a bandpass filter 142, an RMS level
25 detector 144, a fixed gain amplifier 146, a bandpass filter 148, an RMS level
26 detector 150, and a reciprocal generator 152.

27 The difference signal is applied to the input of fixed preemphasis filter 132
28 which generates therefrom an output signal that is applied via line 132d to an
29 input terminal of amplifier 134. An output signal generated by reciprocal
30 generator 152 is applied via line 152a to a gain control terminal of amplifier 134.
31 Amplifier 134 generates an output signal by amplifying the signal on line 132d
32 using a gain that is proportional to the value of the signal on line 152a. The

1 output signal generated by amplifier 134 is applied via line 134a to an input
2 terminal of variable emphasis filter 136, and an output signal generated by RMS
3 detector 144 is applied via line 144a to a control terminal of filter 136. Variable
4 emphasis filter 136 generates an output signal by preemphasizing or
5 deemphasizing the high frequency portions of the signal on line 134a under the
6 control of the signal on line 144a. The output signal generated by filter 136 is
7 applied to the input of overmodulation protector and bandlimiter 138 which
8 generates therefrom the encoded difference signal.

9 The encoded difference signal is applied via feedback path 138a to the
10 inputs of fixed gain amplifiers 140, 146, which amplify the encoded difference
11 signal by Gain A and Gain B, respectively. The amplified signal generated by
12 amplifier 140 is applied to an input of bandpass filter 142 which generates
13 therefrom an output signal that is applied to the input of RMS level detector 144.
14 The latter generates an output signal as a function of the RMS value of the input
15 signal level received from filter 142. The amplified signal generated by amplifier
16 146 is applied to the input of bandpass filter 148 which generates therefrom an
17 output signal that is applied to the input of RMS level detector 150. The latter
18 generates an output signal as a function of the RMS value of the input signal level
19 received from filter 148. The output signal of detector 150 is applied via line
20 150a to reciprocal generator 152, which generates a signal on line 152a that is
21 representative of the reciprocal of the value of the signal on line 150a. As stated
22 above, the output signals generated by RMS level detector 144 and reciprocal
23 generator 152 are applied to filter 136 and amplifier 134, respectively.

24 As shown in Figure 1, the difference channel processing section 130 is
25 considerably more complex than the sum channel processing section 120. The
26 additional processing provided by the difference channel processing section 130,
27 in combination with complementary processing provided by a decoder (not
28 shown) receiving a BTSC signal, maintains the signal-to-noise ratio of the
29 difference channel at acceptable levels even in the presence of the higher noise
30 floor associated with the transmission and reception of the difference channel.
31 Difference channel processing section 130 essentially generates the encoded
32 difference signal by dynamically compressing, or reducing the dynamic range of

1 the difference signal so that the encoded signal may be transmitted through the
2 limited dynamic range transmission path associated with a BTSC signal, and so
3 that a decoder receiving the encoded signal may recover all the dynamic range in
4 the original difference signal by expanding the compressed difference signal in a
5 complementary fashion. The difference channel processing section 130 is a
6 particular form of the adaptive signal weighing system described in U.S. Patent
7 No. 4,539,526, which is known to be advantageous for transmitting a signal
8 having a relatively large dynamic range through a transmission path having a
9 relatively narrow, frequency dependent, dynamic range.

10 Briefly, the difference channel processing section may be thought of as
11 including a wide band compression unit 180 and a spectral compression unit 190.
12 The wide band compression unit 180 includes variable gain amplifier 134
13 preferably in the form of a voltage controlled amplifier, and the components of
14 the feedback path for generating the control signal to amplifier 134 and
15 comprising amplifier 146, band pass filter 148, RMS level detector 150, and
16 reciprocal generator 152. Band pass filter 148 has a relatively wide pass band,
17 weighted towards lower audio frequencies, so in operation the output signal
18 generated by filter 148 and applied to RMS level detector 150 is substantially
19 representative of the encoded difference signal. RMS level detector 150 therefore
20 generates an output signal on line 150a representative of a weighted average of
21 the energy level of the encoded difference signal, and reciprocal generator 152
22 generates a signal on line 152a representative of the reciprocal of this weighted
23 average. The signal on line 152a controls the gain of amplifier 134, and since
24 this gain is inversely proportional to a weighted average (i.e., weighted towards
25 lower audio frequencies) of the energy level of the encoded difference signal,
26 wide band compression unit 180 "compresses", or reduces the dynamic range, of
27 the signal on line 132a by amplifying signals having relatively low amplitudes and
28 attenuating signals having relatively large amplitudes.

29 The spectral compression unit 190 includes variable emphasis filter 136
30 and the components of the feedback path generating a control signal to the filter
31 136 and comprising amplifier 140, band pass filter 142 and RMS level detector
32 144. Unlike filter 148, band pass filter 142 has a relatively narrow pass band

1 that is weighted towards higher audio frequencies. As is well known, the
2 transmission medium associated with the difference portion of the BTSC
3 transmission system has a frequency dependent dynamic range and the pass band
4 of filter 142 is chosen to correspond to the spectral portion of that transmission
5 path having the narrowest dynamic range (i.e., the higher frequency portion). In
6 operation the output signal generated by filter 142 and applied to RMS level
7 detector 144 contains primarily the high frequency portions of the encoded
8 difference signal. RMS level detector 144 therefore generates an output signal on
9 line 144a representative of the energy level in the high frequency portions of the
10 encoded difference signal. This signal then controls the preemphasis/deemphasis
11 applied by variable emphasis filter 136 so in effect the spectral compression unit
12 190 dynamically compresses high frequency portions of the signal on line 134a by
13 an amount determined by the energy level in the high frequency portions of the
14 encoded difference signal as determined by the filter 142. The use of the spectral
15 compression unit 190 thus provides additional signal compression towards the
16 higher frequency portions of the difference signal, which combines with the
17 wideband compression provided by the variable gain amplifier 134 to effectively
18 cause more overall compression to take place at high frequencies relative to the
19 compression at lower frequencies. This is done because the difference signal
20 tends to be noisier in the higher frequency part of the spectrum. When the
21 encoded difference signal is decoded with a wideband expander and a spectral
22 expander in a decoder (not shown), respectively in a complementary manner to
23 the wide band compression unit 180 and spectral compression unit 190 of the
24 encoder, the signal-to-noise ratio of the L-R signal applied to the difference
25 channel processing section 130 will be substantially preserved.

26 The BTSC standard rigorously defines the desired operation of the 75 μ s
27 preemphasis filter 122, the fixed preemphasis filter 132, the variable emphasis
28 filter 136, and the bandpass filters 142, 148, in terms of idealized analog filters.
29 Specifically, the BTSC standard provides a transfer function for each of these
30 components and the transfer functions are described in terms of mathematical
31 representations of idealized analog filters. The BTSC standard also defines the
32 gain settings, Gain A and Gain B, of amplifiers 140 and 146, respectively, and

1 also defines the operation of amplifier 134, RMS level detectors 144, 150, and
2 reciprocal generator 152. The BTSC standard also provides suggested guidelines
3 for the operation of overmodulation protector and bandlimiter 138 and bandlimiter
4 124. Specifically, bandlimiter 124 and the bandlimiter portion of overmodulation
5 protector and bandlimiter 138 are described as low pass filters with cutoff
6 frequencies of 15 kHz, and the overmodulation protection portion of
7 overmodulation protector and bandlimiter 138 is described as a threshold device
8 that limits the amplitude of the encoded difference signal to 100% of full
9 modulation where full modulation is the maximum permissible deviation level for
10 modulating the audio subcarrier in a television signal.

11 Since encoder 100 is defined in terms of mathematical descriptions of
12 idealized filters it may be thought of as an idealized or theoretical encoder, and
13 those skilled in the art will appreciate that it is virtually impossible to construct a
14 physical realization of a BTSC encoder that exactly matches the performance of
15 theoretical encoder 100. Therefore, it is expected that the performance of all
16 BTSC encoders will deviate somewhat from the theoretical ideal, and the BTSC
17 standard defines maximum limits on the acceptable amounts of deviation. For
18 example, the BTSC standard states that a BTSC encoder must provide at least 30
19 db of separation from 100 Hz to 8,000 Hz where separation is a measure of how
20 much a signal applied to only one of the left or right channel's inputs appears
21 erroneously in the other of the left or right channel's outputs.

22 The BTSC standard also defines a composite stereophonic baseband signal
23 (referred to hereinafter as the "composite signal") that is used to generate the
24 audio portion of a BTSC signal. The composite signal is generated using the
25 conditioned sum signal, the encoded difference signal, and a tone signal,
26 commonly referred to as the "pilot tone" or simply as the "pilot", which is a sine
27 wave at a frequency f_H where f_H is equal to 15,734 Hz. The presence of the pilot
28 in a received television signal indicates to the receiver that the television signal is
29 a BTSC signal rather than a monophonic or other non-BTSC signal. The
30 composite signal is generated by multiplying the encoded difference signal by a
31 waveform that oscillates at twice the pilot frequency according to the cosine
32 function $\cos(4\pi f_H t)$, where t is time, to generate an amplitude modulated, double-

1 sideband, suppressed carrier signal and by then adding to this signal the
2 conditioned sum signal and the pilot tone.

3 Figure 2 is a graph of the spectrum of the composite signal. In Figure 2
4 the spectral band of interest containing the content of the conditioned sum signal
5 (or the "sum channel signal") is indicated as "L+R", the two spectral sidebands
6 containing the content of the frequency shifted encoded difference signal (or the
7 "difference channel signal") are each indicated as "L-R", and the pilot tone is
8 indicated by the arrow at frequency f_H . As shown in Figure 2, in the composite
9 signal the encoded difference signal is used at 100% of full modulation, the
10 conditioned sum signal is used at 50% of full modulation, and the pilot tone is
11 used at 10% of full modulation.

12 Stereophonic television has been widely successful, and existing encoders
13 have performed admirably, however, virtually every BTSC encoder now in use
14 has been built using analog circuitry technology. These analog BTSC encoders,
15 and particularly the analog difference channel processing sections, due to their
16 increased complexity have been relatively difficult and expensive to construct.
17 Due to the variability of analog components, complex component selection and
18 extensive calibration have been required to produce acceptable analog difference
19 channel processing sections. Further, the tendency of analog components to drift,
20 over time, away from their calibrated operating points has also made it difficult to
21 produce an analog difference channel processing section that consistently and
22 repeatably performs within a given tolerance. A digital difference channel
23 processing section, if one could be built, would not suffer from these problems of
24 component selection, calibration, and performance drift, and could potentially
25 provide increased performance.

26 Further, the analog nature of existing BTSC encoders has made them
27 inconvenient to use with newly developed, increasingly popular, digital
28 equipment. For example, television programs can now be stored using digital
29 storage media such as a hard disk or digital tape, rather than the traditional
30 analog storage media, and in the future increasing use will be made of digital
31 storage media. Generating a BTSC signal from a digitally stored program now
32 requires converting the digital audio signals to analog signals and then applying

1 the analog signals to an analog BTSC encoder. A digital BTSC encoder, if one
2 could be built, could accept the digital audio signals directly and could therefore
3 be more easily integrated with other digital equipment.

4 While a digital BTSC encoder would potentially offer several advantages,
5 there is no simple way to construct an encoder using digital technology that is
6 functionally equivalent to the idealized encoder 100 defined by the BTSC
7 standard. One problem is that the BTSC standard defines all the critical
8 components of idealized encoder 100 in terms of analog filter transfer functions.
9 As is well known, while it is generally possible to design a digital filter so that
10 either the magnitude or the phase response of the digital filter matches that of an
11 analog filter, it is extremely difficult to match both the amplitude and phase
12 responses without requiring large amounts of processing capacity for processing
13 data sampled at very high sampling rates or without significantly increasing the
14 complexity of the digital filter. Without increasing either the sampling frequency
15 or the filter order, the amplitude response of a digital filter can normally only be
16 made to more closely match that of an analog filter at the expense of increasing
17 the disparity between the phase responses of the two filters, and vice versa.
18 However, since small errors in either amplitude or phase decrease the amount of
19 separation provided by BTSC encoders, it would be essential for a digital BTSC
20 encoder to closely match both the amplitude and phase responses of an idealized
21 encoder of the type shown at 100 in Figure 1.

22 For a digital BTSC encoder to provide acceptable performance, it is
23 critical to preserve the characteristics of the analog filters of an idealized encoder
24 100. Various techniques exist for designing a digital filter to match the
25 performance of an analog filter; however, in general, none of these techniques
26 produce a digital filter (of the same order as the analog filter) having amplitude
27 and phase responses that exactly match the corresponding responses of the analog
28 filter. Ideal encoder 100 is defined in terms of analog transfer functions specified
29 in the frequency domain, or the s-plane, and to design a digital BTSC encoder,
30 these transfer functions must be transformed to the z-plane. Such a
31 transformation may be performed as a "many-to-one" mapping from the s-plane
32 to the z-plane which attempts to preserve time domain characteristics. However,

1 in such a transformation the frequency domain responses are subject to aliasing
2 and may be altered significantly. Alternatively, the transformation may be
3 performed as a "one-to-one" mapping from the s-plane to the z-plane that
4 compresses the entire s-plane into the unit circle of the z-plane. However, such a
5 compression suffers from the familiar "frequency warping" between the analog
6 and digital frequencies. Prewarping can be employed to compensate for this
7 frequency warping effect, however, prewarping does not completely eliminate the
8 deviations from the desired frequency response. These problems would have to
9 be overcome to produce a digital BTSC encoder that performs well and is not
10 unduly complex or expensive.

11 There is therefore a need for overcoming these difficulties and developing
12 a digital BTSC encoder.

13 Objects of the Invention

14 It is an object of the present invention to substantially reduce or overcome
15 the above-identified problems of the prior art.

16 Another object of the present invention is to provide an adaptive digital
17 weighing system.

18 Still another object of the present invention is to provide an adaptive
19 digital weighing system for encoding an electrical information signal of a
20 predetermined bandwidth so that the information signal can be recorded on or
21 transmitted through a dynamically-limited, frequency dependent channel having a
22 narrower dynamically-limited portion in a first spectral region than in at least one
23 other spectral region of the predetermined bandwidth.

24 And another object of the present invention is to provide a digital BTSC
25 encoder.

26 Yet another object of the present invention is to provide a digital BTSC
27 encoder that prevents ticking, a problem that can arise with substantially zero
28 input signal levels.

29 And another object of the present invention is to provide a digital BTSC
30 encoder that uses a sampling frequency that is a multiple of a pilot tone signal
31 frequency of 15,734 Hz so as to prevent interference between the signal
32

1 information of the encoded signal with the pilot tone signal.

2 Still another object of the invention is to provide a digital BTSC encoder
3 for generating a conditioned sum signal and an encoded difference signal that
4 include substantially no signal energy at the pilot tone frequency of 15,734 Hz.

5 Yet another object of the present invention is to provide a digital BTSC
6 encoder including a sum channel processing section for generating the conditioned
7 sum signal, and a difference processing section for generating the encoded
8 difference signal, the sum channel processing section including devices for
9 introducing compensatory phase errors into the conditioned sum signal to
10 compensate for any phase errors introduced into the encoded difference signal by
11 the difference channel processing section.

12 And another object of the present invention is to provide a digital BTSC
13 encoder including a digital variable emphasis unit, the unit including a digital
14 variable emphasis filter characterized by a variable coefficient transfer function,
15 and the unit further including a device for selecting the coefficients of the variable
16 coefficient transfer function as a function of the signal energy of the encoded
17 difference signal.

18 Yet another object of the present invention is to provide a digital BTSC
19 encoder including a composite modulator for generating a composite modulated
20 signal from the conditioned sum signal and the encoded difference signal.

21 Still another object of the present invention is to provide a digital BTSC
22 encoder that may be implemented on a single integrated circuit.

23 24 Summary of the Invention

25 These and other objects are provided by an improved BTSC encoder that
26 includes an input section, a sum channel processing section, and a difference
27 channel processing section all of which are implemented using digital technology.
28 In one aspect, the input section includes high pass filters for preventing the BTSC
29 encoder from exhibiting "ticking". In another aspect, the BTSC encoder uses a
30 sampling frequency that is equal to an integer multiple of the pilot frequency.

31 In yet another aspect, the sum channel processing section generates a
32 conditioned sum signal, and the difference channel processing section generates

1 an encoded difference signal, and the sum channel processing section includes
2 components for introducing a phase error into the conditioned sum signal to
3 compensate for any phase errors introduced into the encoded difference signal by
4 the difference channel processing section.

5 According to yet another aspect, the invention provides an adaptive digital
6 weighing system for encoding an electrical information signal of a predetermined
7 bandwidth so that the information signal can be recorded on or transmitted
8 through a dynamically-limited, frequency dependent channel having a narrower
9 dynamically-limited portion in a first spectral region than in at least one other
10 spectral region of the predetermined bandwidth.

11 Still other objects and advantages of the present invention will become
12 readily apparent to those skilled in the art from the following detailed description
13 wherein several embodiments are shown and described, simply by way of
14 illustration of the best mode of the invention. As will be realized, the invention
15 is capable of other and different embodiments, and its several details are capable
16 of modifications in various respects, all without departing from the invention.
17 Accordingly, the drawings and description are to be regarded as illustrative in
18 nature, and not in a restrictive or limiting sense, with the scope of the application
19 being indicated in the claims.

20 21 Brief Description of the Drawings

22 For a fuller understanding of the nature and objects of the present
23 invention, reference should be had to the following detailed description taken in
24 connection with the accompanying drawings in which the same reference
25 numerals are used to indicate the same or similar parts wherein:

26 Figure 1 shows a block diagram of a prior art idealized BTSC encoder;

27 Figure 2 shows a graph of the spectrum of the composite signal generated
28 in accordance with the BTSC standards;

29 Figure 3 shows a block diagram of one embodiment of a digital BTSC
30 encoder constructed according to the invention;

31 Figures 4A-C show block diagrams of low pass filters used in the digital
32 BTSC encoder shown in Figure 3;

1 Figure 5 shows a detailed block diagram of the wideband compression unit
2 used in the digital BTSC encoder shown in Figure 3;

3 Figure 6 shows a block diagram of the spectral compression unit used in
4 the digital BTSC encoder shown in Figure 3;

5 Figure 7 shows a flow chart used for calculating the filter coefficients of
6 the variable emphasis filter used in the spectral compression unit shown in Figure
7 6;

8 Figures 8A-D show block diagrams that illustrate signal scaling that may
9 be used to preserve resolution and decrease the chance of saturation in fixed point
10 implementations of digital BTSC encoders constructed according to the invention;

11 Figure 9 shows a detailed block diagram of the composite modulator
12 shown in Figures 8B-C; and

13 Figure 10 shows a block diagram of one preferred embodiment of sum and
14 difference channel processing sections that may be used in digital BTSC encoders
15 constructed according to the invention.

16

17 Detailed Description of the Drawings

18 Figure 3 is a block diagram of one embodiment of a digital BTSC encoder
19 200 constructed according to the invention. Digital encoder 200 is constructed to
20 provide performance that is functionally equivalent to the performance of
21 idealized encoder 100 (shown in Figure 1). As with idealized encoder 100,
22 digital encoder 200 receives the left and right channel audio input signals and
23 generates therefrom the conditioned sum signal and the encoded difference signal,
24 however, in digital encoder 200 these input and output signals are digitally
25 sampled signals rather than continuous analog signals.

26 The choice of sampling frequency f_s for the left and right channel audio
27 input signals significantly affects the design of digital encoder 200. In the
28 preferred embodiments, the sampling frequency f_s is chosen to be an integer
29 multiple of the pilot frequency f_H , so that $f_s = Nf_H$ where N is an integer, and in
30 the most preferred embodiments, N is selected to be greater than or equal to
31 three. It is important for encoder 200 to insure that the conditioned sum and
32 encoded difference signals do not contain enough energy at the pilot frequency f_H

1 to interfere with the pilot tone that is included in the composite signal. As will
2 be discussed in greater detail below, it is therefore desirable for at least some of
3 the filters in digital encoder 200 to provide an exceptionally large degree of
4 attenuation at the pilot frequency f_H , and this choice of sampling frequency f_s
5 simplifies the design of such filters.

6 Digital encoder 200 includes an input section 210, a sum channel
7 processing section 220 and a difference channel processing section 230. Rather
8 than simply implementing the difference channel processing section 230 using
9 digital technology, all three sections 210, 220, 230 are implemented entirely using
10 digital technology. Many of the individual components in digital encoder 200
11 respectively correspond to individual components in idealized encoder 100. In
12 general, the components of digital encoder 200 have been selected so that their
13 amplitude responses closely match the respective amplitude responses of their
14 corresponding components in encoder 100. This often results in there being a
15 relatively large difference between the phase responses of corresponding
16 components. According to one aspect of the present invention, means are
17 provided in digital encoder 200 for compensating for or nullifying these phase
18 differences, or phase errors. As those skilled in the art will appreciate, relatively
19 small phase errors in the difference channel processing section 230 may be
20 compensated for by introducing similar phase errors in the sum channel
21 processing section 220, and implementing the sum channel processing section
22 using digital technology simplifies the introduction of such desired compensating
23 phase errors.

24 The input section 210 of encoder 200 includes two high pass filters 212,
25 214, and two signal adders 216, 218. The left channel digital audio input signal
26 L is applied to the input of high pass filter 212, the latter generating therefrom an
27 output signal that is applied to positive input terminals of adders 216, 218. The
28 right channel audio input signal R is applied to the input of high pass filter 214
29 which generates therefrom an output signal that is applied to a positive input
30 terminal of adder 216 and to a negative input terminal of adder 218. Adder 216
31 generates a sum signal (indicated in Figure 3 as "L+R") by summing the output
32 signals generated by filters 212 and 214. Adder 218 generates a difference signal

1 (indicated in Figure 3 as "L-R") by subtracting the output signal generated by
2 filter 214 from the output signal generated by filter 212. Input section 210 is
3 therefore similar to input section 110 (shown in Figure 1) however, section 210
4 additionally includes the two high pass filters 212, 214 and generates digital sum
5 and difference signals.

6 High pass filters 212, 214 preferably have substantially identical responses
7 and preferably remove D.C. components from the left and right channel audio
8 input signals. As will be discussed in greater detail below, this D.C. removal
9 prevents encoder 200 from exhibiting a behavior referred to as "ticking". Since
10 the audio information content of the left and right channel audio input signals of
11 interest is considered to be within a frequency band between 50 Hz and 15,000
12 Hz, removal of D.C. components does not interfere with the transmission of the
13 information content of the audio signals. Filters 212, 214, therefore, preferably
14 have a cutoff frequency below 50 Hz, and more preferably have a cutoff
15 frequency below 10 Hz so that they will not remove any audio information
16 contained in the audio input signals. Filters 212, 214 also preferably have a flat
17 magnitude response in their passband. In one preferred embodiment, filters 212,
18 214 are implemented as first order infinite impulse response (IIR) filters, each
19 having a transfer function $H(z)$ given by the formula shown in the following
20 Equation (1).

21
$$H(z) = \frac{1 - z^{-1}}{1 + a_1 z^{-1}} \quad (1)$$

22 Referring again to Figure 3, the sum channel processing section 220
23 receives the sum signal and generates therefrom the conditioned sum signal. In
24 particular, the sum signal is applied to a 75 μ s preemphasis filter 222. The filter
25 222 in turn generates an output signal that is applied to a static phase
26 equalization filter 228. The filter 228 generates an output signal that is applied to
27 a low pass filter 224 of section 220 which in turn generates the conditioned sum
28 signal.

29 The 75 μ s preemphasis filter 222 provides signal processing that is
30 partially analogous to the filter 122 (shown in Figure 1) of idealized encoder 100.

1 The amplitude response of filter 222 is preferably selected to closely match that
 2 of filter 122. As will be discussed further below, means are preferably provided
 3 in difference channel processing section 230 for compensation for any differences
 4 in the phase responses of filters 222 and 122. In one preferred embodiment,
 5 filter 222 is implemented as a first order IIR filter having a transfer function $H(z)$
 6 that is described by the formula shown in the following Equation (2).

$$7 \quad H(z) = \frac{b_0 + b_1 z^{-1}}{1 + a_1 z^{-1}} \quad (2)$$

8 Static phase equalization filter 228 performs processing that is not directly
 9 analogous to any of the components in idealized encoder 100 (shown in Figure 1).
 10 As will be discussed in greater detail below, static phase equalization filter 228 is
 11 used to introduce phase errors that compensate for phase errors introduced by
 12 difference processing section 230. Briefly, static phase equalization filter 228 is
 13 preferably an "all-pass" filter having a relatively flat amplitude response and a
 14 selected phase response. In one preferred embodiment, filter 228 is implemented
 15 as a first order IIR filter having a transfer function $H(z)$ that is described by the
 16 formula shown in the following Equation (3).

$$17 \quad H(z) = \frac{a_0 + z^{-1}}{1 + a_0 z^{-1}} \quad (3)$$

18 Low pass filter 224 provides processing that is partially analogous to
 19 bandlimiter 124 (shown in Figure 1) of encoder 100. Low pass filter 224
 20 preferably provides a flat amplitude response in a pass band of zero to 15 kHz
 21 and a relatively sharp cutoff above 15kHz. Filter 224 also preferably provides an
 22 exceptionally large degree of attenuation at the frequency f_H of the pilot tone (i.e.,
 23 15,734 Hz). By providing this exceptionally large degree of attenuation, filter
 24 224 insures that the conditioned sum signal does not include enough energy at the
 25 pilot frequency f_H to interfere with the pilot tone used in the composite signal.
 26 As discussed above, selecting the sampling frequency f_s to be equal to an integer
 27 multiple of the pilot frequency f_H simplifies the design of a filter that provides an
 28 exceptionally large degree of attenuation at the pilot frequency and therefore
 29 simplifies the design of filter 224. Filter 224 preferably has a null at the pilot

1 frequency f_H and preferably provides at least 70 dB of attenuation for all
 2 frequencies from the pilot frequency f_H up to one-half the sample rate.

3 Figure 4A is a block diagram illustrating one preferred embodiment of low
 4 pass filter 224. As shown in Figure 4A, filter 224 may be implemented by
 5 cascading five filter sections 310, 312, 314, 316, 318. In one preferred
 6 embodiment, all five filter sections 310, 312, 314, 316, 318 are each
 7 implemented as a second order IIR filter having transfer functions $H(z)$ which are
 8 described by the formula shown in the following Equation (4).

$$9 \quad H(z) = \frac{b_0 + b_1 z^{-1} + b_2 z^{-2}}{1 + a_1 z^{-1} + a_2 z^{-2}} \quad (4)$$

10 So in the embodiment shown in Fig. 4A, filter 224 is tenth order IIR filter.

11 Referring again to Figure 3, the difference channel processing section 230
 12 receives the difference signal and generates therefrom the encoded difference
 13 signal. The difference signal is applied to a low pass filter 238a which generates
 14 therefrom an output signal that is applied to a fixed preemphasis filter 232a. The
 15 latter generates an output signal that is applied via line 239 to an input terminal of
 16 a wideband compression unit 280, and the encoded difference signal is applied via
 17 feedback line 240 to a detector terminal of wideband compression unit 280. The
 18 latter generates an output signal that is applied via line 281 to an input terminal of
 19 a spectral compression unit 290, and the encoded difference signal is also applied
 20 via feedback line 240 to a detector terminal of unit 290. The latter generates an
 21 output signal that is applied to a fixed preemphasis filter 232b which in turn
 22 generates an output signal that is applied to a clipper 254. Clipper 254 generates
 23 an output signal that is applied to a low pass filter 238b which in turn generates
 24 the encoded difference signal.

25 Low pass filters 238a, 238b, together form a low pass filter 238 that
 26 performs processing that is partially analogous to the bandlimiter portion of
 27 overmodulation protector and bandlimiter 138 (shown in Figure 1) of idealized
 28 encoder 100. Preferably, filter 238 is implemented so that it is substantially
 29 identical to low pass filter 224, which is used in the sum channel processing
 30 section 220. Any phase errors introduced into the encoded difference signal by

1 filter 238 are therefore compensated by balancing phase errors that are introduced
2 into the conditioned sum signal by filter 224. Filter 238 is preferably split into
3 two sections 238a, 238b as shown for reasons which will be discussed in greater
4 detail below, and filter 238a preferably has a null at the pilot frequency f_H .

5 Figures 4B-C are block diagrams illustrating one preferred embodiment of
6 the respective filters 238a and 238b. As shown in Figure 4B, filter 238a may be
7 implemented by cascading three filter sections 310, 314, 318 that are identical to
8 three of the filter sections used in filter 224 (shown in Figure 4A), and as shown
9 in Figure 4C, filter 238b may be implemented by cascading two filter sections
10 312, 316 that are identical to the two remaining sections used in filter 224.

11 Fixed preemphasis filters 232a, 232b (shown in Figure 3) together form a
12 fixed preemphasis filter 232 that performs processing that is partially analogous to
13 filter 132 (shown in Figure 1) of idealized encoder 100. The amplitude response
14 of filter 232 is preferably selected to closely match the amplitude response of
15 filter 132. In one embodiment, the phase responses of filters 232 and 132 are
16 significantly different, and as will be discussed in greater detail below, the
17 resulting phase errors are compensated for by filters 222 and 228 in the sum
18 channel processing section 220. Filter 232 is preferably split into two sections
19 232a, 232b as shown for reasons that will be discussed below. In one preferred
20 embodiment, filters 232a, 232b are each implemented as first order IIR filters
21 having transfer functions $H(z)$ that are described by the formula shown in
22 Equation (2). So in this embodiment filter 232 is a second order IIR filter.

23 In one preferred embodiment, the difference between the phase responses
24 of filters 232b and 132a closely matches the difference between the phase
25 responses of filters 222 and 122. Therefore, the phase error introduced into the
26 encoded difference signal by fixed preemphasis filter 232b is balanced by the
27 phase error introduced into the conditioned sum signal by 75 μ s preemphasis filter
28 222. Further, in this embodiment, the phase response of static phase equalization
29 filter 228 is selected to closely match the difference between the phase responses
30 of fixed preemphasis filter 232a and filter 132b, so that any phase error
31 introduced into the encoded difference signal by filter 232a is balanced by a
32 compensatory phase error in the conditioned sum signal that is introduced by

1 static phase equalization filter 228. .

2 Clipper 254 performs processing that is partially analogous to the
3 overmodulation protection portion of overmodulation protector and bandlimiter
4 138 (shown in Figure 1) used in idealized encoder 100. Briefly, clipper 254 is
5 implemented as a thresholding device, however, the operation of clipper 254 will
6 be discussed in greater detail below.

7 Wideband compression unit 280 and spectral compression unit 290
8 perform processing functions that are partially analogous to that of units 180 and
9 190, respectively, of idealized encoder 100 (shown in Figure 1). Briefly,
10 wideband compression unit 280 dynamically compresses the signal on line 239 as
11 a function of the overall energy level in the encoded difference signal and spectral
12 compression unit 290 further compresses high frequency portions of the signal on
13 line 281 as a function of high frequency energy in the encoded difference signal.

14 Figure 5 shows a block diagram of a preferred embodiment of a digital
15 wideband compression unit 280. Unit 280 includes a digital signal multiplier
16 434, a digital signal multiplier 446, a wideband digital bandpass filter 448, a
17 digital RMS level detector 450, and a digital reciprocal generator 458. These
18 components perform processing functions partially analogous to those performed
19 by amplifier 134, amplifier 146, bandpass filter 148, RMS level detector 150, and
20 reciprocal generator 152, respectively, of idealized encoder 100 (shown in Figure
21 1). The encoded difference signal is applied via feedback path 240 to an input of
22 wideband digital bandpass filter 448 which generates therefrom an output signal
23 that is applied to RMS level detector 450. The latter generates an output signal
24 that is representative of the RMS value of the output signal generated by filter
25 448 and applies this output signal via line 450a to reciprocal generator 458.
26 Reciprocal generator 458 then generates an output signal representative of the
27 reciprocal of the signal on line 450a and applies this output signal via line 458a to
28 multiplier 446. Digital signal multiplier 446 multiplies the signal on line 458a by
29 the value of the gain setting, Gain D, and thereby generates an output signal that
30 is representative of D times the reciprocal of the RMS value and that is applied
31 via line 446a to an input terminal of multiplier 434. The output signal generated
32 by fixed preemphasis filter 232a is applied via line 239 to another input terminal

1 of multiplier 434. Multiplier 434 multiplies the signal on line 239 by the signal
2 on line 446a and thereby generates the output of wideband compression unit 280
3 which is applied via line 281 to the input of spectral compression unit 290.

4 Wideband digital bandpass filter 448 is designed to have an amplitude
5 response that closely matches the amplitude response of bandpass filter 148
6 (shown in Figure 1). One preferred choice is to select filter 448 so that the mean
7 square difference between its amplitude response and that of filter 148 are
8 minimized. In one embodiment, the phase response of filters 448 and 148 are
9 substantially different, but since the output signal of the RMS level detector 450
10 is substantially insensitive to the phase of its input signal, these phase differences
11 may be ignored. In one preferred embodiment, wideband bandpass filter 448 is
12 implemented as a second order IIR filter having a transfer function $H(z)$ that is
13 described by the formula shown in Equation (4).

14 RMS level detector 450 is designed to approximate the performance of
15 detector 150 which is used in idealized encoder 100 (shown in Figure 1).
16 Detector 450 includes a signal squaring device 452, a signal averaging device
17 454, and a square root device 456. Squaring device 452 squares the signal
18 generated by bandpass filter 448 and applies this squared signal via line 452a to
19 averaging device 454. The latter computes a time weighted average of the signal
20 on line 452a and applies the average via line 454a to square root device 456.
21 Square root device 456 calculates the square root of the signal on line 454a and
22 thereby generates a signal on line 450a representative of the RMS value of the
23 output signal generated by wideband digital bandpass filter 448.

24 Averaging device 454 includes a digital signal multiplier 460, a digital
25 signal adder 462, a digital signal multiplier 464, and a delay register 465. The
26 output signal generated by squaring device 452 is applied via line 452a to one
27 input of multiplier 460 which generates an output signal by scaling the signal on
28 line 452a by a constant α . The scaled output signal generated by multiplier 460
29 is applied to one input of adder 462 and an output signal generated by delay
30 register 465 is applied to the other input of adder 462. Adder 462 generates an
31 output signal by summing the signals present at its two inputs, and this summed
32 signal is the output signal of averaging device 454 and is applied to square root

1 device 456 via line 454a. This summed signal is also applied to one input of
2 multiplier 464 which generates an output signal by scaling the summed signal by
3 the constant $(1-\alpha)$. The output signal generated by multiplier 464 is applied to
4 an input of delay register 465. Those skilled in the art will appreciate that
5 averager 454 is a recursive filter and implements a digital averaging function that
6 is described by the recursive formula shown in the following Equation (5).

$$y(n)=\alpha x(n)+(1-\alpha)y(n-1) \quad (5)$$

7
8
9
10 in which $y(n)$ represents the current digital sample of the signal output by
11 averager 454 on line 454a, $y(n-1)$ represents the previous digital sample of the
12 signal output by averager 454 on line 454a, and $x(n)$ represents the current digital
13 sample of the signal output by squaring device 452 on line 452a. Those skilled in
14 the art will appreciate that averager 454 provides a digital approximation of the
15 analog averaging function defined in the BTSC standard and implemented by
16 RMS level detector 150 (shown in Figure 1) of idealized encoder 100. The
17 constant α is preferably chosen so that the time constant of RMS level detector
18 450 closely approximates the corresponding time constant specified in the BTSC
19 standard for RMS level detector 150.

20 Digital square root device 456 and digital reciprocal generator 458 are
21 shown in Figure 5 as two separate components, however, those skilled in the art
22 will appreciate that these two components may be implemented using a single
23 device that generates an output signal representative of the reciprocal of the
24 square root of its input signal. Such a device may be implemented for example
25 as a memory look up table (LUT), or alternatively may be implemented using
26 processing components that calculate a Taylor series polynomial approximation of
27 the inverse square root function.

28 Figure 6 shows a block diagram of a preferred embodiment of spectral
29 compression unit 290. Unit 290 includes a variable preemphasis/deemphasis unit
30 (hereinafter referred to as the "variable emphasis unit") 536, a signal multiplier
31 540, a spectral band pass filter 542, and an RMS level detector 544, and these
32 components provide processing which is partially analogous to that of variable

1 emphasis filter 136, amplifier 140, bandpass filter 142, and RMS level detector
2 144, respectively, of idealized encoder 100 (shown in Figure 1). The encoded
3 difference signal is applied via feedback line 240 to an input of signal multiplier
4 540 which generates an output signal by multiplying the encoded difference signal
5 by the fixed gain setting value of Gain C. The amplified output signal generated
6 by signal multiplier 540 is applied to spectral bandpass filter 542 which generates
7 an output signal that is applied to RMS level detector 544. The latter generates
8 an output signal that is applied via line 544a to a control terminal of variable
9 emphasis unit 536, and the output signal generated by wideband compressor unit
10 280 is applied via line 281 to an input terminal of unit 536. The latter
11 dynamically varies the frequency response applied to the signal on line 281
12 according to a function of the signal on line 544a, the latter signal being a
13 function of the signal energy of the encoded difference signal within the
14 frequency band passed by spectral band pass filter 542. The output signal of unit
15 290, which is generated by unit 536 and is applied to the input of fixed
16 preemphasis filter 232b, is thus dynamically compressed a greater amount in the
17 high frequency portions of the signal than in the remainder of the spectrum of
18 interest.

19 Spectral bandpass filter 542 is designed to have an amplitude response that
20 closely matches the amplitude response of bandpass filter 142 (shown in Figure 1)
21 of idealized encoder 100. As with filter 448 (shown in Figure 5), one preferred
22 choice is to select filter 542 so that the difference between its RMS amplitude
23 response and that of filter 142 are minimized. In one embodiment, the phase
24 response of filters 542 and 142 are substantially different, but since the RMS
25 output of RMS level detector 544 is substantially insensitive to the phase of the
26 input to the detector, these phase differences may be ignored. In one preferred
27 embodiment, spectral bandpass filter 542 is implemented as a cascade of three
28 second order IIR filter sections 542a, 542b, 542c (as shown in Figure 6) each
29 having a transfer function $H(z)$ that is described by the formula shown in
30 Equation (4):

31 RMS level detector 544 is designed to approximate the performance of
32 detector 144 which is used in idealized encoder 100 (shown in Figure 1).

1 Detector 544 includes a signal squaring device 552, a signal averaging device
2 554, and a square root device 556. Squaring device 552 squares the signal
3 generated by spectral bandpass filter 542 and applies this squared signal via line
4 552a to averaging device 554. The latter functions similarly to averaging device
5 454 (shown in Figure 5) which is used in the wideband compression unit 280,
6 although device 554 preferably uses a constant β different from the constant α .
7 The behavior of averaging device 554 is of course also described by Equation (5)
8 when β is substituted for α . The constant β is preferably selected for device 554
9 so that the time constant of RMS level detector 544 closely approximates the
10 corresponding time constant specified by the BTSC standard for RMS level
11 detector 144 (shown in Figure 1). Averaging device 554 computes a time
12 weighted average of the signal on line 552a and applies the average to square root
13 device 556 via line 554a. Square root device 556 calculates the square root of the
14 signal on line 554a and thereby generates a signal on line 544a as a function of
15 the RMS value of the output signal generated by spectral bandpass filter 542.

16 The signal on line 544a is applied to the control terminal of variable
17 emphasis unit 536. Variable emphasis unit 536 performs processing that is
18 partially analogous to filter 136 (shown in Figure 1) of idealized encoder 100. As
19 defined by the BTSC standard, filter 136 has amplitude and phase responses that
20 vary as a function of the output signal generated by RMS level detector 144. One
21 preferred way to implement unit 536 so that it has similar variable responses is to
22 use a digital filter having variable coefficients that determine its transfer function
23 and to select the value of the coefficients during any given sample period, or
24 group of sample periods, based on the value of the signal on line 544a.

25 Figure 6 shows one embodiment of variable emphasis unit 536 which
26 includes a logarithmic generator 558, a variable emphasis filter 560, and a look
27 up table LUT 562. The output signal generated by RMS level detector 544 is
28 applied via line 544a to logarithmic generator 558. The latter generates a signal
29 on line 558a that is representative of the logarithm of the signal on line 544a and
30 applies this signal to LUT 562. LUT 562 generates an output signal selected
31 from the LUT and representative of filter coefficients to be used by variable
32 emphasis filter 560. The coefficients thus generated by LUT 562 are applied via

1 line 562a to a coefficient selection terminal of variable emphasis filter 560. The
2 output signal generated by wideband compression unit 280 is applied to an input
3 terminal of variable emphasis filter 560 via line 281. Variable emphasis filter
4 560 generates the output signal of spectral compression unit 290 which is applied
5 to the input of fixed preemphasis filter 232b.

6 Variable emphasis filter 560 is designed to have a variable amplitude
7 response that closely matches the variable amplitude response of filter 136 (shown
8 in Figure 1) of idealized encoder 100. Variable emphasis filter 560 provides a
9 similar variable response by using a variable coefficient transfer function (i.e., the
10 coefficients of the transfer function $H(z)$ of filter 560 are variable) and by
11 allowing LUT 562 to select the value of the coefficients during intervals based on
12 the sample period. As will be described in greater detail below, LUT 562 stores
13 the values of the filter coefficients used by filter 560, and during each sample
14 period, or during any selected group of sample periods, LUT 562 selects a set of
15 filter coefficients as a function of the output signal generated by logarithmic
16 generator 558 on line 558a. In one preferred embodiment, variable emphasis
17 filter 560 is implemented as a first order IIR filter having a transfer function $H(z)$
18 that is described by the formula shown in the following Equation (6).

$$19 \quad H(z) = \frac{b_0 + b_1 z^{-1}}{1 + a_1 z^{-1}} \quad (6)$$

20 in which the filter coefficients b_0 , b_1 , and a_1 are variables that are selected by
21 LUT 562. Methods of selecting the values for the filter coefficients used by filter
22 560 as well as by the other filters of encoder 200 will be discussed below.

23 In Figure 6, logarithmic generator 558 and square root device 556 are
24 shown, for convenience, as two separate components. However, those skilled in
25 the art will appreciate that these two components may be implemented using a
26 single device, such as a LUT, or alternatively using processing components that
27 calculate a Tayler series polynomial approximation of the logarithm of the signal
28 on line 554a and by then dividing this value by two. Similarly, in alternative
29 implementations, the functions performed by logarithmic generator 558, square
30 root device 556, and LUT 562 may be incorporated into a single device.

1 As stated above, high pass filters 212, 214 (shown in Figure 3) are useful
2 in blocking DC components so as to prevent encoder 200 from exhibiting a
3 behavior known as "ticking". In the context of a stereophonic encoder, ticking
4 refers to relatively low frequency oscillatory behavior of the encoder caused when
5 there is no signal present at the left and right channel audio inputs. The desired
6 behavior of a stereophonic system when there is no signal present at the audio
7 inputs is to remain silent; however, an encoder connected through a decoder to
8 loudspeakers and exhibiting ticking causes the loudspeakers to emit an audible
9 sound, referred to as a "tick", with a somewhat regular period that is partially
10 dependent on the time constant of the RMS level detector in the wideband
11 compressor. More particularly, in encoder 200, when only very low level signals
12 are present at the audio inputs, and when there is a D.C. component, or an
13 offset, present in the signal on line 239, wideband compression unit 280 tends to
14 behave in an unstable fashion that causes ticking.

15 Consider the case where only a low level audio signal is present on line
16 239. In such a case, the output of RMS level detector 450 on line 450a becomes
17 very small, which in turn causes the gain of multiplier 434 to become very large.
18 If such a low level audio signal on line 239 is constant in its amplitude, the
19 wideband compression unit 280 reaches a steady-state condition after some time
20 (determined by the time constant α applied to multiplier 460), because the
21 encoded difference signal is fed back on line 240 to the wideband compression
22 unit 280. Because the feedback is arranged to be negative, when the audio signal
23 on line 239 increases in its amplitude, the signal on line 450a increases, which in
24 turn causes the gain of multiplier 434 to decrease. When the audio signal on line
25 239 decreases in its amplitude, the signal on line 450a decreases, which in turn
26 causes the gain of multiplier 434 to increase.

27 However, should there be a significant dc signal present on line 239 in
28 addition to a low level audio signal, the dc signal is blocked from the feedback
29 process by the action of wideband bandpass filter 448, which has zero response to
30 dc signals. In particular, any dc present in the encoded difference signal at line
31 240 is blocked by filter 448, and is not sensed by RMS level detector 450. Any
32 dc signal present on line 239 will be amplified by multiplier 434 along with any

1 audio signal present on line 239, but the amplification factor or gain will be
2 determined only by the audio signal amplitude as sensed by RMS level detector
3 450 after filtering by filter 448.

4 As noted above, whenever the amplitude of the audio signal on line 239
5 varies, the gain of multiplier 434 varies inversely. During such variations in
6 gain, any dc present on line 239 will also be subjected to variable amplification,
7 in effect modulating the dc signal, thereby producing an ac signal. In this fashion
8 such dc signals may be modulated so as to create significant audio-band signals
9 which will not be rejected by filter 448, and are therefore sensed by detector 450.
10 When the audio signal on line 239 is small compared to the dc on line 239, small
11 variations in the audio signal level, which cause changes in the gain of amplifier
12 434, can cause a large change in the dc level (which amount to an ac signal) at
13 line 281 through this modulation process. The ac signal produced tends to
14 increase the overall signal which passes through filter 448, regardless of whether
15 the audio signal variation that gave rise to the ac signal was an increase or
16 decrease in signal level. In particular, should the level of the audio signal on line
17 239 decrease, the negative feedback process normally increases the gain of
18 multiplier 434. However, if a sufficient dc signal is present in line 239, a
19 decrease in audio signal on line 239 can cause an increase in the signal sensed by
20 detector 450, forcing the gain of multiplier 434 to decrease. In this fashion, the
21 negative feedback process is reversed, and the feedback becomes positive.

22 Such positive feedback will only persist so long as the modulated dc signal
23 at line 281 is sufficiently large compared to any audio signal present on line 281,
24 when weighted by the response of all the filters and signal modifiers between line
25 281 and the output of filter 448. Once the gain of multiplier 434 decreases
26 sufficiently such that the modulated dc signal in line 281 no longer provides a
27 significant input to detector 450, the feedback reverts to its normal negative
28 sense. In accordance with the time constant of detector 450, the system will re-
29 acquire an appropriate gain level based on the level of the audio signal in line
30 239. But, if sufficient dc remains in the signal in line 239, the cycle will repeat
31 itself once the gain of multiplier 434 increases sufficiently. During each such
32 period of positive feedback, a sharp change in the dc level of line 281 is

1 produced. This change is audible, and sounds somewhat similar to the 'tick' of a
2 clock. Since such dc changes will occur with some regularity, based on the time
3 constant of detector 450, the phenomenon is often referred to as 'ticking'.

4 One method of preventing ticking is to remove any dc components present
5 in the input signal to encoder 200. This is accomplished by high pass filters 212
6 and 214. Further, high pass filters 212 and 214 help to maximize the dynamic
7 range of encoder 200 by removing dc components which otherwise may use up
8 valuable dynamic range. As stated above and as shown in Figure 3, low pass
9 filter 238 is preferably implemented as two filters 238a and 238b. Splitting filter
10 238 in this fashion provides several advantages. If filter 238a were eliminated,
11 and the entire filter 238 were located after clipper 254 (i.e., in the location of
12 filter 238b) then any components above 15 kHz on the audio input signals may
13 cause instability in the wideband compression unit 280 similar to the above-
14 described ticking behavior. This occurs because any signal components above 15
15 kHz on line 239 will be amplified by multiplier 434 (shown in Figure 5) and
16 because such components will not be sensed by RMS level detector 450 since
17 such components are filtered out by the low pass filter following clipper 254
18 (shown in Figure 3). Since detector 450 increases the gain of multiplier 434
19 when it senses the absence of a signal, the gain of multiplier 434 can become
20 relatively large when the signal on line 239 consists of little audio signal (under
21 15kHz) information, but significant high-frequency (over 15kHz) information.
22 Multiplier 434 then amplifies the high-frequency information, which can generate
23 large signals that are likely to be clipped by components in processing section
24 230. This clipping can produce harmonics which may alias to low frequencies
25 that will be sensed by RMS level detector 450 causing the system to tick as
26 described previously. Alternatively, if filter 238b were eliminated and the entire
27 filter 238 were located before fixed preemphasis filter 232a (i.e., in the location
28 of filter 238a) then high frequency artifacts generated by clipper 254 would be
29 included in the encoded difference signal and could interfere with the pilot tone in
30 the composite signal. Therefore, splitting filter 238 as shown provides an optimal
31 arrangement whereby filter 238a prevents ticking in compression unit 280 and
32 filter 238b filters high frequency artifacts that may be generated by clipper 254.

1 Fixed preemphasis filter 232 is also preferably split into two filters 232a,
 2 232b as shown in Figure 3. Filter 232 typically requires relatively large gain at
 3 high frequencies, as is specified in the BTSC standard, and using only a single
 4 section to implement filter 232 increases the likelihood of filter 232 causing
 5 clipping. It is advantageous to apply some of the gain of filter 232 on the input
 6 side of wideband compression unit 280 (with filter 232a) and to apply some of the
 7 gain of filter 232 on the output side of wideband compression unit 280 (with filter
 8 232b). Since unit 280 normally compresses its input signal, distributing the gain
 9 of filter 232 around the compression provided by unit 280 decreases that the
 10 likelihood that the gain of filter 232 will cause an overflow condition.

11 To minimize size, power consumption, and cost, encoder 200 is preferably
 12 implemented using a single digital signal processing chip. Encoder 200 has been
 13 successfully implemented using one of the well known Motorola DSP 56002
 14 digital signal processing chips (this implementation shall be referred to hereinafter
 15 as the "DSP Embodiment"). The Motorola DSP 56002 is a fixed point twenty-
 16 four bit chip, however, other types of processing chips, such as floating point
 17 chips, or fixed point chips having other word lengths, could of course be used.
 18 The DSP Embodiment of encoder 200, uses a sampling frequency f_s that is equal
 19 to three times the pilot frequency f_H (i.e., $f_s = 47202$ Hz). The following Table
 20 1 lists all of the filter coefficients used in the DSP Embodiment of encoder 200
 21 except those used in variable emphasis filter 560.

TABLE 1

23	Low Pass Filter (Section #1) 310	Low Pass Filter (Section #2) 312
24	(Equation 4)	(Equation 4)
25	$b_0=0.18783270$	$b_0=0.44892888$
26	$b_1=0.36310206$	$b_1=0.70268024$
27	$b_2=0.18783270$	$b_2=0.44892888$
28	$a_1=-0.388832539$	$a_1=0.12638618$
29	$a_2=0.12709286$	$a_2=0.47415181$
30	Low Pass Filter (Section #3) 314	Low Pass Filter (Section #4) 316
31	(Equation 4)	(Equation 4)

1	$b_0=0.70674027$	$b_0=0.85733126$
2	$b_1=0.87637648$	$b_1=0.91505047$
3	$b_2=0.70674027$	$b_2=0.85733126$
4	$a_1=0.53702472$	$a_1=0.74320197$
5	$a_2=0.75298490$	$a_2=0.89832289$
6	Low Pass Filter (Section #5) 318	Wideband Bandpass Filter 448
7	(Equation 4)	(Equation 4)
8	$b_0=0.92737972$	$b_0=-0.02854672$
9	$b_1=0.92729649$	$b_1=-0.18789051$
10	$b_2=0.92737972$	$b_2=0.21643723$
11	$a_1=0.82951974$	$a_1=-1.75073141$
12	$a_2=0.97259237$	$a_2=0.75188028$
13	Fixed Preemphasis Filter 238a	Fixed Preemphasis Filter 238b
14	(Equation 2)	(Equation 2)
15	$b_0=9.50682180$	$b_0=4.357528$
16	$b_1=-9.00385663$	$b_1=-3.24843271$
17	$a_1=-0.497064357$	$a_1=0.10881833$
18	Spectral Bandpass Filter	Spectral Bandpass Filter
19	(Section #1) 542a	(Section #2) 542b
20	(Equation 4)	(Equation 4)
21	$b_0=0.646517841$	$b_0=0.850281278$
22	$b_1=0.649137616$	$b_1=-0.850247036$
23	$b_2=0.0$	$b_2=0.0$
24	$a_1=0.557821757$	$a_1=-0.602159890$
25	$a_2=0.0$	$a_2=0.0$
26	Spectral Bandpass Filter	Static Phase Equalization Filter 224
27	(Section #3) 542c	(Equation 3)
28	(Equation 4)	
29	$b_0=0.597678418$	$a_0=0.9029$

1	$b_1 = -1.195357770$	
2	$b_2 = 0.597679348$	
3	$a_1 = -0.776566094$	
4	$a_2 = 0.352824276$	
5	75 μ s preemphasis filter 222	High Pass Filters 212, 214
6	(Equation 2)	(Equation 1)
7	$b_0 = 4.57030583$	$a_1 = -0.999$
8	$b_1 = -3.43823487$	
9	$a_1 = 0.131778883$	

11 In the DSP Embodiment of encoder 200 the value of the constant α that is
 12 used by averager 454 (shown in Figure 5) in wideband compression unit 280 is
 13 set equal to 0.0006093973517, and the value of the constant β that is used by
 14 averager 554 (shown in Figure 6) in spectral compression unit 290 is set equal to
 15 0.001825967. Further, the values of Gain C and Gain D used by amplifiers 540
 16 and 446, respectively, in the spectral and wideband compression units are set
 17 equal to 0.5011872 and 0.08984625, respectively, to insure that the DSP
 18 Embodiment of encoder 200 performs similarly to encoder 100.

19 Figure 7 shows a flow chart 700 that describes one preferred method for
 20 pre-calculating all the sets of filter coefficients used by variable emphasis filter
 21 560 (shown in Figure 6) in the DSP Embodiment of encoder 200. Prior to
 22 operation of encoder 200, all the sets of filter coefficients used by filter 560 are
 23 pre-calculated (e.g., by a general purpose digital computer) and are loaded into
 24 LUT 562. In the DSP Embodiment of encoder 200, filter 560 has a transfer
 25 function $H(z)$ that is described by Equation (6) so flow chart 700 describes the
 26 calculation of the coefficients b_0 , b_1 , and a_1 . As specified in the BTSC standard,
 27 the transfer function of $S(f,b)$ of analog filter 136 (shown in Figure 1) to which
 28 filter 560 partially corresponds, is described by the formula shown in the
 29 following Equation (7).

$$S(f, b) = \frac{1 + \frac{\left(\frac{if}{F}\right)^{(b+51)}}{(b+1)}}{1 + \frac{\left(\frac{if}{F}\right)^{(1+51b)}}{(b+1)}} \quad (7)$$

in which F is equal to 20.1 kHz.

The first step in flow chart 700 in an initialization step 710 during which several variables are initialized. Specifically, the sampling frequency f_s is set equal to 47202 Hz, and the period T is set equal to $1/f_s$. The variable W is a digital version of the variable F used in Equation (7) and is set equal to $\pi(20.1 \text{ kHz})/f_s$. The variable dBRANGE represents the desired signal range of the RMS detectors in the spectral compression unit, and for the DSP Embodiment dBRANGE is set equal to 72.25 dB. The variable dBRES relates to the sensitivity of filter 560 to changes in the energy level of the encoded difference signal. In the DSP Embodiment of encoder 200, dBRES is set equal to 0.094 dB so that filter 560 will use coefficients based on the value of the signal on line 558a quantized to the nearest 0.094 dB. The variable N equals the total number of sets of filter coefficients used in filter 560 and N is calculated by dividing the sensitivity (dBRES) into the range (dBRANGE) and rounding to the nearest integer. In the DSP Embodiment, N is equal to 768 although those skilled in the art will appreciate that this number can be changed which will vary the sensitivity or the range. In the DSP Embodiment, LUT 562 stores 769 sets of coefficients for filter 560, and of course if N is increased, a larger LUT will be used to store the extra sets of filter coefficients. Further, those skilled in the art will appreciate that logarithmic generator 558 scales the signal on line 558a and thereby reduces the number of filter coefficient sets stored by LUT 562, for a given minimum quantization of the value of the signal on line 558a. However, in other embodiments, logarithmic generator 558 may be eliminated and LUT 562 may store a correspondingly larger number of filter coefficient sets. Finally, the variables Scale and Address are set equal to 32 and zero, respectively. The

1 variable Scale, which is only used in fixed point implementations, is selected so
2 that all the filter coefficients have a value greater than or equal to negative one
3 and less than one (where the filter coefficients are represented in twos
4 complement).

5 Following initialization step 710, a coefficient generation step 720 is
6 executed. During the first execution of step 720, variables $b_0(0)$, $b_1(0)$, and $a_1(0)$
7 are calculated which correspond to values of the coefficients b_0 , b_1 , and a_1 that
8 are to be stored at address location zero of LUT 562. Following this execution of
9 step 720, an incrementing step 730 is executed during which the value of the
10 variable Address is incremented. Following step 730 a comparison step is
11 executed during which the values of the variables Address and N are compared.
12 If Address is less than or equal to N, then steps 720, 730, and 740 are reexecuted
13 iteratively so that values of the coefficients b_0 , b_1 , and a_1 are calculated for each
14 of the 769 addresses of LUT 562. When step 740 detects that the value of
15 Address is greater than N, then all 769 sets of coefficients have been calculated
16 and execution of flow chart 700 proceeds to a concluding step 750.

17 In coefficient generation step 720, the variable dBFS corresponds to the
18 output of logarithmic generator 558. As the value of the variable Address ranges
19 from zero to 769, the value of dBFS ranges from about -72.25 to zero dB
20 corresponding to the signal range of about 72.25 dB provided by the DSP
21 Embodiment of encoder 200 (where zero dB corresponds to the full modulation).
22 The variable RMSd corresponds to the output of the analog RMS level detector
23 144 (shown in Figure 1), and as the variable Address ranges from zero to 769,
24 the value of RMSd ranges from about -36 to 36 dB corresponding to the signal
25 range of 72 dB provided by typical prior art analog BTSC encoders. The
26 variable RMSb is a linear version of the variable RMSd, and RMSb corresponds
27 to the variable b in the transfer function $S(f,b)$ described in Equation (7). The
28 variables K1 and K2 correspond to the $(b+51)/(b+1)$ and the $(51b+1)/(b+1)$
29 terms, respectively, in Equation (7). The coefficients b_0 , b_1 , and a_1 are
30 calculated as shown in step 720 using the variables K1, K2, W, and Scale.

31 Figure 8A shows a block diagram that illustrates one method of using the
32 DSP Embodiment in an analog system, and in Figure 8A, all components that are

1 implemented in the 56002 integrated circuit are indicated at 200a. The analog
2 system supplies analog left and right channel audio input signals (shown in Figure
3 8A as "L" and "R", respectively) and these signals are applied to the inputs of
4 sixteen bit analog-to-digital converters 810 and 812, respectively. Converters
5 810, 812 sample their analog input signals using a sampling frequency f_s that is
6 equal to 47,202 Hz (i.e., $3f_H$) and converters 810, 812 thereby generate sequences
7 of sixteen bit digital samples that are representative of the left and right channel
8 audio input signals, respectively. The signals generated by converters 810 and
9 812 are applied to encoder 200a where they are received by modules 292 and
10 294, respectively. Modules 292, 294 are "divide by sixteen" modules (which
11 divide the amplitude of their inputs by a factor of 16) and therefore generate
12 output signals that are equal to their input signals divided by sixteen. Since
13 division by any power of two is easily accomplished in a digital system by using
14 a shift register, modules 292, 294 are implemented as shift registers that shift
15 their inputs by four binary places.

16 As stated above, the 56002 chip is a fixed point twenty-four bit processor,
17 and the samples applied to the chip by converters 810, 812 are in twos
18 complement representation. Modules 292, 294 divide the samples generated by
19 converters 810, 812 by sixteen and thereby place each of the samples in the
20 middle of a twenty-four bit word. So in every sample generated by modules 292,
21 294, the four most significant bits are sign bits and the four least significant bits
22 are zeros, and the sixteen bits in the middle of the word correspond to one
23 sample generated by one of the converters 810, 812. Padding each twenty-four
24 bit word with sign bits at the high end and with zeros at the low end in this
25 fashion preserves accuracy and allows intermediate signals generated by encoder
26 200a to exceed sixteen bits without causing an error condition such as an
27 overflow.

28 In encoder 200a, each bit of the twenty-four bit word corresponds roughly
29 to 6dB of signal range, and therefore modules 292, 294 correspond to -24 dB
30 (i.e., negative 6 times 4) attenuators. If the analog input signals applied to
31 converters 810, 812 are considered for reference purposes as zero dB signals,
32 then the signals generated by modules 292, 294 are attenuated by 24 dB.

1 Input section 210 receives the twenty-four bit words generated by modules
2 292, 294 and generates therefrom the sum signal that is applied to the sum
3 channel processing section 220. The output signal generated by sum channel
4 processing section 220 is applied to a "times 16 module" (which may be
5 considered as a 24 dB amplifier) 296. Module 296 thereby compensates for the -
6 24dB attenuators 292, 294 and brings the output of sum channel processing
7 section 220 back to 100% modulation (i.e., back to "full scale"). The output
8 signal generated by module 296 is applied to a sixteen bit digital-to-analog
9 converter 814 which in turn generates an analog conditioned sum signal.

10 Input section 210 also generates the difference signal that is applied to the
11 difference channel processing section 230. As stated above, as a result of
12 modules 292, 294, the difference signal may be considered as being attenuated by
13 24 dB. In the DSP embodiment of encoder 200a, clipper 254 (shown in Figure
14 3) of the difference processing section 230 includes an 18 dB amplifier (which is
15 implemented as a multiply by eight). That is, clipper 254 amplifies the signal
16 generated by fixed preemphasis filter 232b by 18 dB and then clips this amplified
17 signal so that the output signal generated by clipper 254 will not exceed a number
18 that is 6dB down from full modulation. The signal applied from clipper 254 to
19 low pass filter 238b therefore has one bit (or 6 dB) of "headroom", so filter 238b
20 may generate an output signal that is 6 dB greater than its input signal without
21 causing saturation. It is desirable to leave this one bit of headroom because the
22 transient response of filter 238b includes some ringing that may cause it to
23 temporarily generate an instantaneous output signal that is greater than its
24 instantaneous input signal and the headroom thereby prevents any ringing in filter
25 238b from causing a saturation condition. Referring again to Figure 8A, the
26 output signal generated by filter 238b is applied to a sixteen bit digital-to-analog
27 converter 816 which in turn generates an output signal that is applied to a 6 dB
28 analog amplifier 820. Both D/A converters 814 and 816 are intended to be
29 complete converters, which include the well-known analog anti-image filters as
30 part of their functionality. Anti-image filters are analog filters applied to the
31 analog signal following digital to analog conversion which serve to attenuate any
32 images of the desired signal which are mirrored about the sample frequency and

1 multiples thereof. Converters 814 and 816 are assumed to be substantially
2 identical to one another, running at the same sample rate and containing
3 substantially the same anti-image filtering. Such converters are commonly
4 available in commercial embodiments, such as the Crystal Semiconductor
5 CS4328. Amplifier 820 amplifies its input signal by 6 dB and thereby brings the
6 encoded difference signal back up to full scale. While Figure 8A shows encoder
7 200a coupled to analog-to-digital converters 810, 812 for receiving analog audio
8 signals, in digital systems converters 810, 812 may of course be eliminated so
9 that encoder 200a receives the digital audio signals directly.

10 Figure 8B shows a block diagram of one preferred embodiment of a BTSC
11 encoder 200b constructed according to the invention and configured as part of an
12 analog system. Encoder 200b is similar to encoder 200a, however, in encoder
13 200b module 296 amplifies its input signal by 18 dB (by multiplying by 8) rather
14 than by 24 dB as in encoder 200a. The output signal generated by module 296 is
15 a scaled version of the conditioned sum signal and is shown in Figure 8B as S.
16 Also, encoder 200b includes a module 298 for amplifying the output signal
17 generated by difference channel processing section 230 by 6 dB (by multiplying
18 by two). The output signal generated by module 298 is a scaled version of the
19 encoded difference signal and is shown in Figure 8B as D. Further, encoder
20 200b includes a composite modulator 822 for receiving the signals S and D and
21 for generating therefrom a digital version of the composite signal. The digital
22 composite signal generated by modulator 822 is applied to a digital-to-analog
23 converter 818 the output of which is an analog version of the composite signal.
24 D/A converter 818 is intended to be a complete converter which includes the
25 aforementioned analog anti-image filter as part of its functionality. Such
26 converters are commonly available in commercial embodiments, such as the Burr-
27 Brown PCM1710. In the preferred embodiments, modules 292, 294, input
28 section 210, sum channel processing section 220, difference channel processing
29 section 230, modules 296, 298, and composite modulator 822 are all implemented
30 on a single digital signal processing chip.

31 Since the composite signal is generated as a digital signal in encoder 200b,
32 module 298 is included to bring the output signal generated by difference channel

1 processing section 230 up to full scale rather than waiting until after digital-to-
2 analog conversion and using an analog amplifier such as amplifier 820 as is
3 shown in Figure 8A. Also, since in the composite signal the conditioned sum
4 signal is used at 50% modulation, module 296 only amplifies its input signal by
5 18 dB so that the output signal generated by module 296 is at half the amplitude
6 of the output signal generated by module 298.

7 Figure 9 shows a block diagram of one embodiment of composite
8 modulator 822. The latter receives the signals S and D and generates therefrom a
9 digital version of the composite signal. Modulator 822 includes two interpolators
10 910, 912, two digital low pass filters 914, 916, a digital signal multiplier 918,
11 and two digital signal adders 920, 922. The S and D signals are applied to
12 respective inputs of the interpolators 910 and 912. Interpolators 910, 912, which
13 are alternatively referred to as "up-samplers", interpolate a new sample between
14 every two consecutive samples applied to their inputs, and thereby generate
15 output signals having twice the sampling frequency as the input signals S and D.
16 The output signals generated by interpolators 910 and 912 are applied to
17 respective inputs of low pass filters 914 and 916. The latter remove images
18 introduced into the S and D signals by interpolators 910, 912. The filtered output
19 signal generated by filter 916 is applied to one input of signal multiplier 918 and
20 a digital oscillating signal as a function of $\cos[4\pi(f_H/f_S)n]$ is applied to the other
21 input of multiplier 918. Multiplier 918 thereby generates the amplitude
22 modulated, double-sideband, suppressed carrier version of the difference signal
23 that is used in the composite signal. The output signal generated by multiplier
24 918 is applied to one input of signal adder 920 and the filtered output signal
25 generated by filter 914 is applied to the other input of signal adder 920. The
26 latter generates an output signal by summing the two signals present at its inputs
27 and applies this signal to signal adder 922. A pilot tone signal that oscillates as a
28 function of $A\cos[2\pi(f_H/f_S)n]$ (where 'A' is a constant representative of 10% of
29 full scale modulation) is applied to the other input of signal adder 922 which
30 generates the digital composite signal by summing the two signals present at its
31 inputs.

32 Composite modulator 822 includes interpolators 910, 912 because the

1 highest frequency component in the composite signal is slightly less than $3f_H$ (as
2 is shown in Figure 2), and therefore the signals applied to the inputs of signal
3 multiplier 918 and signal adder 920 should have sample rates at least as large as
4 $6f_H$ to satisfy the Nyquist criteria. Because the sample rate at the output of
5 composite modulator 822 is typically higher than the sample rate of either the S
6 or D signals, D/A converter 818 must be capable of operating at such higher
7 sample rates. If the input signals S and D applied to composite modulator 822
8 have sample rates of $3f_H$ some form of interpolation (such as that provided by
9 interpolators 910, 912) should be provided to double the sample rate. Of course,
10 if sufficiently high sample rates are used throughout encoder 200b then
11 interpolators 910, 912 and low pass filters 914, 916 may be eliminated from
12 modulator 822.

13 Figure 8C shows a block diagram of yet another embodiment of a BTSC
14 encoder 200c constructed according to the invention. Encoder 200c is similar to
15 encoder 200b (shown in Figure 8B) however, in encoder 200c module 298 is
16 eliminated so that the signal generated by the difference channel processing
17 section 230 is the signal D and is applied directly to composite modulator 822.
18 Further, in encoder 200c, module 296 amplifies its input signal by 12 dB (by
19 multiplying by 4) rather than by 18 dB as is done in encoder 200b. So in
20 encoder 200c, the signals S and D are 6 dB down from the levels of those signals
21 in encoder 200b. Composite modulator 822 therefore generates from these
22 signals a version of the composite signal that is attenuated by 6 dB. This
23 attenuated version of the composite signal is converted to an analog signal by
24 digital-to-analog converter 818 and is then brought up to full scale by 6 dB analog
25 amplifier 820. As with encoder 200b, encoder 200c is preferably implemented
26 using a single digital signal processing chip.

27 The differences between encoders 200b and 200c represent design
28 tradeoffs. As those skilled in the art will appreciate, when converting a digital
29 signal to an analog signal with a digital-to-analog converter, insuring that the
30 digital signal is at full scale tends to minimize any loss of signal-to-noise ratio
31 that might occur as a result of the conversion. Encoder 200b minimizes the loss
32 of signal-to-noise ratio as a result of the operation of converter 818 by using

1 modules 296, 298 to insure that the digital version of the composite signal
2 (generated by modulator 822) that is applied to converter 818 is at full scale.
3 However, although converter 200b minimizes any loss of signal-to-noise ratio that
4 might occur as a result of converter 818, encoder 200b also increases the
5 likelihood that clipping might occur in the composite signal. Since the difference
6 channel processing section 230 uses the relatively large gain provided by fixed
7 preemphasis filter 232 (shown in Figure 3), it is possible for some clipping to
8 occur in the path of the encoded difference signal. Encoder 200b uses module
9 298 to bring the D signal up to full scale and this essentially eliminates any
10 headroom from the signal path of the D signal and thereby increases the chance
11 that some clipping will occur. So encoder 200b minimizes the loss of any signal-
12 to-noise ratio that occurs as a result of converter 818 at the cost of increasing the
13 likelihood of clipping in the path of the encoded difference signal. In contrast,
14 encoder 200c preserves headroom in the path of the encoded difference signal and
15 thereby reduces the likelihood of clipping at the cost of increasing the loss of
16 signal-to-noise ratio that occurs as a result of operation of converter 818.

17 Figure 8D shows a block diagram of yet another embodiment of a BTSC
18 encoder 200d constructed according to the invention. Encoder 200d is similar to
19 encoder 200a (shown in Figure 8A) however, encoder 200d additionally includes
20 a portion 822a of a composite modulator. Portion 822a includes two interpolators
21 910, 912, two low pass filters 914, 916, digital signal multiplier 918 and a digital
22 signal adder 930. The S signal generated by module 296 is applied to
23 interpolator 910 which "up-samples" the S signal and applies the up-sampled
24 signal to low pass filter 914. The latter filters this signal and applies the filtered
25 signal to one input terminal of adder 930. A digital pilot tone having twice the
26 normal amplitude (i.e., $2A\cos 2\pi(f_H/f_S)n$) is applied to the other input terminal of
27 adder 930 which generates an output signal by summing the two signals present at
28 its input terminals. The D signal generated by difference channel processing
29 section 230 is applied to interpolator 912 which generates an up-sampled signal
30 that is applied to low pass filter 916. The latter filters this signal and applies the
31 filtered signal to one terminal of multiplier 918. A signal oscillating according to
32 $\cos 4\pi(f_H/f_S)n$ is applied to the other terminal of multiplier 918 which generates an

1 output signal by multiplying the two signals present at its input terminals. As
2 with encoders 200a-c, encoder 200d is preferably implemented using a single
3 digital signal processing chip.

4 Encoder 200d is preferably used in conjunction with two digital-to-analog
5 converters 932, 934, an analog -6 dB attenuator 936, an analog 6 dB amplifier
6 938, and an analog adder 940. The output signal generated by adder 930 is
7 applied to converter 932 which generates an analog signal that is applied to
8 attenuator 936. The output signal generated by multiplier 918 is applied to
9 converter 934 which generates an analog signal that is applied to amplifier 938.
10 The signals generated by attenuator 936 and amplifier 938 are applied to input
11 terminals of signal adder 940 which sums these signals to generate the analog
12 composite signal. D/A converters 932 and 934 are intended to be complete
13 converters which include the aforementioned analog anti-image filters as part of
14 their functionality. Converters 932 and 934 are assumed to be substantially
15 identical to one another, running at the same sample rate and containing
16 substantially the same anti-image filtering. Such converters are commonly
17 available in commercial embodiments, such as the Burr Brown PCM1710.

18 It is also possible to eliminate interpolator 910 and low pass filter 914
19 from Figure 8D, and run D/A converter 932 at a sample rate equal to that of the
20 sum channel processing section 220. However, to do so is generally not practical
21 because inexpensive, commonly available D/A converters are usually available in
22 pairs housed within a single integrated circuit. Such paired D/A converters
23 naturally operate at the same sample rate. While it is possible to reduce DSP
24 complexity by eliminating interpolator 910 and low pass filter 914 from Figure
25 8D, doing so would also likely increase the cost and complexity of the overall
26 design because a simple stereo D/A converter could no longer be used for both
27 D/A converters 932 and 934.

28 Encoder 200d represents one combination of the features of encoders 200b
29 and 200c. Encoder 200d uses module 296 to bring the S signal up to full scale so
30 as to minimize any loss of signal-to-noise ratio that might occur as a result of the
31 operation of converter 932. Encoder 200d also preserves 6 dB of headroom in
32 the signal path of the D signal and therefore reduces the likelihood of any loss of

1 accuracy due to clipping. Although encoder 200d includes more components than
2 either of encoders 200b and 200c, encoder 200d both minimizes loss of signal-to-
3 noise ratio and the likelihood of clipping.

4 Figure 10 shows a block diagram of a preferred embodiment of sum
5 channel processing section 220a and difference channel processing section 230a
6 for use in encoder 200 (and these sections 220a, 230a may of course be used in
7 encoders 200a-d). Processing sections 220a, 230a are similar to the above-
8 described sections 220, 230, however, section 220a additionally includes dynamic
9 phase equalization filter 1010, and section 230a additionally includes a dynamic
10 phase equalization filter 1012. In the illustrated embodiment, the output signals
11 generated by static phase equalization filter 228 and fixed preemphasis filter 232a
12 are applied to the input terminals of dynamic phase equalization filters 1010 and
13 1012, respectively, and the output signal generated by logarithmic generator 558
14 on line 558a is applied to the control terminals of filters 1010, 1012. The output
15 signals generated by filters 1010 and 1012 are applied to low pass filter 224 and
16 to wideband compression unit 280, respectively.

17 Dynamic phase equalization filters 1010, 1012 are used to compensate for
18 phase errors introduced by variable emphasis filter 560 which is used in spectral
19 compression unit 290. The phase response of variable emphasis filter 560 is
20 preferably matched as closely as is possible to that of variable emphasis filter 136
21 (shown in Figure 1). However, due to the variable, signal dependent, nature of
22 variable emphasis filter 136, it is extremely difficult to design variable emphasis
23 filter 560 so that its phase response is matched to that of variable emphasis filter
24 136 for all pre-emphasis/de-emphasis characteristics, which in turn varies with
25 signal level. Therefore in typical embodiments of encoder 200, the phase
26 responses of variable emphasis filter 560 and variable emphasis filter 136 diverge
27 as a function of the signal level. Dynamic phase equalization filters 1010, 1012
28 preferably introduce compensatory phase errors into the sum and difference
29 channel processing sections to compensate for the divergence between variable
30 emphasis filter 560 and variable emphasis filter 136.

31 Dynamic phase equalization filters 1010, 1012 therefore perform a
32 function that is similar to that performed by static phase equalization filter 228.

1 However, whereas filter 228 compensates for phase errors that are independent of
2 the level of the encoded difference signal, filters 1010, 1012 compensate for
3 phase errors that are dependent on this signal level. Filters 1010, 1012 are
4 preferably implemented as "all pass" filters having relatively flat magnitude
5 responses and selected phase responses. Dynamic phase equalization filters are
6 included in both the sum and difference processing sections because a phase delay
7 may be required in either the sum or difference channel to compensate for the
8 phase error introduced by variable emphasis filter 560. In preferred
9 embodiments, filters 1010, 1012 are implemented in a similar fashion as variable
10 emphasis unit 536 and include a filter having a variable coefficient transfer
11 function and a LUT for selecting the values of the filter coefficients during any
12 particular interval. The signal generated by logarithmic generator 558 on line
13 558a is preferably applied to the control terminals of filters 1010, 1012 and
14 selects the filter coefficients used by those filters.

15 Digital encoder 200 has been discussed in connection with certain
16 particular embodiments, however, those skilled in the art will appreciate that
17 variations of these embodiments are also embraced within the invention. For
18 example, variable emphasis unit 536 (shown in Figure 6) has been discussed in
19 terms of being implemented using a variable emphasis filter 560 and a LUT 562.
20 However, rather than precomputing all the possible coefficients for filter 560 and
21 storing them in LUT 562, it may be preferable for other implementations of
22 variable emphasis unit 536 to eliminate LUT 562 and to instead include
23 components for calculating the filter coefficients in real time. Those skilled in
24 the art will appreciate that such considerations represent a tradeoff between
25 memory resources (such as are used by a LUT for storing filter coefficients) and
26 computing resources (such as are used by components for calculating filter
27 coefficients in real time) and may be resolved differently in any particular
28 implementation of encoder 200. Similar considerations apply to square root
29 devices 456 and 556, reciprocal generator 458, and logarithmic generator 558
30 (shown in Figures 5 and 6) which may alternatively use memory resources (e.g.,
31 a LUT for storing all the values) or processing resources (e.g., for calculating a
32 Taylor series polynomial approximation). In yet other embodiments, any or all

1 of the components in encoder 200 may be implemented using individual hardware
2 components or alternatively as software modules running on a general or specific
3 purpose computer.

4 Another example of variations of encoder 200 that are embraced within the
5 invention relates to scaling modules 292, 294 (shown in Figure 8B). These
6 modules are particularly relevant to fixed point implementations of encoder 200.
7 In floating point implementations there is no need to pad each sample with zeros
8 and sign bits to prevent overflow and these modules can therefore be eliminated
9 from floating point implementations. As a further example, the static phase
10 equalization filter 228 (shown in Figure 10) has been discussed in terms of
11 compensating for phase errors introduced by filter 232a, however, filter 228 may
12 be alternatively used to compensate for other phase errors introduced by other
13 components in the difference channel processing section 230a. Still further,
14 filters 228 and 1010 may be implemented as a single filter.

15 Therefore, since certain changes may be made in the above apparatus
16 without departing from the scope of the invention herein involved, it is intended
17 that all matter contained in the above description or shown in the accompanying
18 drawing shall be interpreted in an illustrative and not a limiting sense.

**What is Claimed is:**

- 1 1. A digital BTSC encoder, comprising:
- 2 (A) left high pass filter means for receiving a digital left channel audio signal
- 3 and for digitally high pass filtering said digital left channel audio signal and
- 4 thereby generating a digital left filtered signal;
- 5 (B) right high pass filter means for receiving a digital right channel audio
- 6 signal and for digitally high pass filtering said digital right channel audio signal
- 7 and thereby generating a digital right filtered signal;
- 8 (C) matrix means for receiving said digital left and digital right filtered
- 9 signals, and including means for summing said digital left and digital right
- 10 filtered signals and thereby generating a digital sum signal, and including means
- 11 for subtracting one of said digital left and digital right filtered signals from the
- 12 other of said digital left and digital right filtered signals and thereby generating a
- 13 digital difference signal;
- 14 (D) difference channel processing means for digitally processing said digital
- 15 difference signal;
- 16 (E) sum channel processing means for digitally processing said digital sum
- 17 signal.
- 1 2. An encoder according to claim 1, wherein said left and right high pass
- 2 filter means are characterized by a cutoff frequency that is less than or equal to
- 3 50 Hz.
- 1 3. An encoder according to claim 2, wherein said left and right high pass
- 2 filter means are characterized by a passband and a substantially flat response in
- 3 said passband.

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REPLACED

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- 1 4. A digital BTSC encoder, comprising:
- 2 (A) left channel sampling means for receiving an analog left channel audio
- 3 signal and for sampling said analog left channel audio signal at a sampling
- 4 frequency substantially equal to N times 15,734 Hz and thereby generating a
- 5 digital left signal, N being an integer greater than or equal to three;
- 6 (B) right channel sampling means for receiving an analog right channel audio
- 7 signal and for sampling said analog right channel audio signal at a sampling
- 8 frequency substantially equal to N times 15,734 Hz and thereby generating a
- 9 digital right signal;
- 10 (C) matrix means for receiving said digital left and digital right signals, and
- 11 including means for summing said digital left and digital right signals and thereby
- 12 generating a digital sum signal, and including means for subtracting one of said
- 13 digital left and digital right signals from the other of said digital left and digital
- 14 right signals and thereby generating a digital difference signal;
- 15 (D) difference channel processing means for digitally processing said digital
- 16 difference signal;
- 17 (E) sum channel processing means for digitally processing said digital sum
- 18 signal.

- 1 5. An encoder according to claim 3, further including digital high pass filter
- 2 means for high pass filtering said digital left and right signals.

- 1 6. A digital BTSC encoder, comprising:
- 2 (A) matrix means for receiving a digital left channel audio signal and a digital
- 3 right channel audio signal, and including means for summing said digital left and
- 4 right channel audio signals and thereby generating a digital sum signal, and
- 5 including means for subtracting one of said digital left and right channel audio
- 6 signals from the other of said digital left and right channel audio signals and
- 7 thereby generating a digital difference signal;
- 8 (B) difference channel processing means for digitally processing said digital
- 9 difference signal, said digital processing introducing a first phase error to said
- 10 digital difference signal;





11 (C) sum channel processing means for processing said digital sum signal and
12 including means for introducing a second phase error to said digital sum signal
13 and thereby compensating for said first phase error introduced into said digital
14 difference signal.

1 7. An adaptive digital signal weighing system: including a signal path for
2 transmitting an electrical information signal of a predetermined bandwidth through
3 said system, said system further comprising:

4 digital filter means disposed in said signal path for varying the gain
5 impressed on the portion of said information signal within a first select spectral
6 region within said predetermined bandwidth by a first variable gain factor, said
7 first variable gain factor varying in response to and as a function of a first control
8 signal;

9 means for digitally generating said first control signal only in response to
10 and in accordance with the signal energy of said information signal within a
11 second select spectral region including at least a part of said first select spectral
12 region;

13 digital gain control means disposed in said signal path and coupled to said
14 digital filter means for varying the signal gain impressed on said information
15 signal substantially throughout said predetermined bandwidth by a second variable
16 gain factor, said second variable gain factor varying in response to and as a
17 function of a second control signal; and

18 means for digitally generating said second control signal in response to
19 and as a function of the signal energy of said information signal substantially
20 within a third select spectral region within said predetermined bandwidth.

1 8. A digital system for encoding an electrical information signal of a
2 predetermined bandwidth so that said information signal can be recorded on or
3 transmitted through a dynamically-limited, frequency dependent channel having a
4 narrower dynamically-limited portion in a first spectral region than in at least one
5 other spectral region of said predetermined bandwidth, said system comprising:
6 input means for receiving said information signal;





7 a signal transmission path coupled to said input means for transmitting said
8 information signal received at said input means;
9 output means coupled to said input means through said signal transmission
10 path for providing said information signal as encoded by said system;
11 digital gain control means coupled to said signal path for varying the
12 signal gain impressed on said information signal substantially throughout said
13 predetermined bandwidth, said signal gain varying in response to and as a
14 function of a first control signal;
15 digital filter means coupled to said signal path and said digital gain control
16 means for impressing a second variable gain on the portion of said information
17 signal substantially within said first spectral region so as to preemphasize said
18 portion with respect to the remaining portions of said information signal, said
19 second variable gain varying in response to and as a function of a second control
20 signal;
21 means for digitally generating said first control signal in response to and
22 as a function of the signal energy of said information signal substantially within a
23 second spectral region of said predetermined bandwidth; and
24 means for digitally generating said second control signal only in response
25 to and in accordance with the signal energy of said information signal within a
26 third spectral region of said predetermined bandwidth including at least a part of
27 said first spectral region.

1 9. An adaptive digital signal weighing system, comprising:
2 a signal path for transmitting an electrical information signal of a
3 predetermined bandwidth through said system;
4 variable coefficient digital filter means for filtering said information
5 signal, said filtering being characterized by a variable coefficient transfer function
6 and said filtering varying the gain impressed on the portion of said information
7 signal within a first select spectral region within said predetermined bandwidth by
8 a first variable gain factor, the variable coefficients of said variable coefficient
9 transfer function and said first variable gain factor varying in response to and as a
10 function of a first control signal;





11 means for digitally generating said first control signal only in response to
12 and in accordance with the signal energy of said information signal within a
13 second select spectral region including at least a part of said first select spectral
14 region;
15 digital gain control means disposed in said signal path and coupled to said
16 variable coefficient digital filter means for varying the signal gain impressed on
17 said information signal substantially throughout said predetermined bandwidth by
18 a second variable gain factor, said second variable gain factor varying in response
19 to and as a function of a second control signal; and
20 means for digitally generating said second control signal in response to
21 and as a function of the signal energy of said information signal substantially
22 within a third select spectral region within said predetermined bandwidth.

1 10. A digital BTSC encoder according to claim 1, wherein said sum channel
2 processing means and said difference channel processing means are both
3 implemented on a single integrated circuit.

1 11. A digital BTSC encoder according to claim 1, wherein said digital left and
2 right channel audio signals are digitally sampled signals sampled with a sampling
3 frequency substantially equal to N times 15,734 Hz, N being an integer greater
4 than or equal to three.

1 12. A digital BTSC encoder according to claim 1, wherein said difference
2 channel processing means includes:

- 3 (A) difference input means for receiving said digital difference signal;
4 (B) difference output means for providing an encoded difference signal;
5 (C) a difference signal transmission path coupled to said difference
6 input means and to said difference output means, and including means for
7 generating said encoded difference signal from said digital difference signal.

8 13. A digital BTSC encoder according to claim 12, wherein said difference
9 channel processing means includes spectral compression means for receiving a





10 spectral compression input signal from said difference signal transmission path
11 and for compressing said spectral compression input signal according to a
12 function of an energy level of said encoded difference signal and thereby
13 generating a spectral compression output signal, and including means for applying
14 said spectral compression output signal to said difference signal transmission path.

1 14. A digital BTSC encoder according to claim 13, wherein said spectral
2 compression means includes means for measuring a first energy level of said
3 encoded difference signal in a first select spectral portion, and for generating a
4 first control signal representative of said first energy level.

1 15. A digital BTSC encoder according to claim 14, wherein said spectral
2 compression means includes a variable emphasis filter for receiving and digitally
3 filtering said spectral compression input signal and thereby generating said
4 spectral compression output signal, the filtering provided by said variable
5 emphasis filter being characterized by a transfer function including a plurality of
6 coefficients, said spectral compression means further including means for
7 selecting said plurality of coefficients according to a function of said first control
8 signal.

1 16. A digital BTSC encoder according to claim 15, wherein said means for
2 selecting said plurality of coefficients includes a memory look up table.

1 17. A digital BTSC encoder according to claim 16, wherein said means for
2 selecting said plurality of coefficients includes logarithmic generator means for
3 receiving and logarithmically compressing said first control signal and thereby
4 generating a logarithmically compressed signal and further includes means for
5 applying said logarithmically compressed signal to said memory look up table.

6 18. A digital BTSC encoder according to claim 15, wherein said means for
7 selecting said plurality of coefficients includes means for calculating said plurality
8 of coefficients as a function of said first control signal.





9 19. A digital BTSC encoder according to claim 15, wherein said spectral
10 compression means includes spectral bandpass filter means for receiving and
11 filtering a signal representative of said encoded difference signal and thereby
12 generating a spectral signal, said filtering provided by said spectral bandpass filter
13 means being characterized by a passband in said first select spectral portion.

1 20. A digital BTSC encoder according to claim 19, wherein said spectral
2 compression means includes first RMS level detector means for receiving said
3 spectral signal and for generating therefrom said first control signal, said first
4 control signal being representative of an RMS value of said spectral signal.

1 21. A digital BTSC encoder according to claim 19, wherein said spectral
2 compression means includes an amplifying means for receiving and amplifying
3 said encoded difference signal and thereby generating said signal representative of
4 said encoded difference signal.

1 22. A digital BTSC encoder according to claim 15, wherein said difference
2 channel processing means includes wideband compression means for receiving a
3 wideband compression input signal from said difference signal transmission path
4 and for compressing said wideband compression input signal according to a
5 function of an energy level of said encoded difference signal and thereby
6 generating a wideband compression output signal, and including means for
7 applying said wideband compression output signal to said difference signal
8 transmission path.

1 23. A digital BTSC encoder according to claim 22, wherein said spectral
2 compression input signal comprises said wideband compression output signal.

1 24. A digital BTSC encoder according to claim 22, wherein said wideband
2 compression means includes means for measuring a second energy level of said
3 encoded difference signal in a second select spectral portion, and for generating a
4 second control signal representative of said second energy level.





5 25. A digital BTSC encoder according to claim 24, wherein said wideband
6 compression means includes amplifier means for receiving and amplifying said
7 wideband compression input signal using a gain controlled by said second control
8 signal and thereby generating said wideband compression output signal.

1 26. A digital BTSC encoder according to claim 25, wherein said wideband
2 compression means includes wideband bandpass filter means for receiving and
3 filtering a signal representative of said encoded difference signal and thereby
4 generating a wideband signal, said filtering provided by said wideband bandpass
5 filter means being characterized by a passband in said second select spectral
6 portion.

1 27. A digital BTSC encoder according to claim 26, wherein said wideband
2 compression means includes second RMS level detector means for receiving said
3 wideband signal and for generating therefrom said second control signal, said
4 second control signal being representative of an RMS value of said wideband
5 signal.

1 28. A digital BTSC encoder according to claim 22, wherein said difference
2 channel processing means includes first low pass filter means for receiving a first
3 low pass filter input signal from said difference signal transmission path and
4 including means for low pass filtering said first low pass filter input signal and
5 thereby generating a first low pass filter output signal, and including means for
6 applying said first low pass filter output signal to said difference signal
7 transmission path.

8 29. A digital BTSC encoder according to claim 28, wherein said difference
9 channel processing means includes second low pass filter means for receiving a
10 second low pass filter input signal from said difference signal transmission path
11 and including means for low pass filtering said second low pass filter input signal
12 and thereby generating a second low pass filter output signal, and including
13 means for applying said second low pass filter output signal to said difference





14 signal transmission path.

1 30. A digital BTSC encoder according to claim 29, wherein said sum channel
2 processing means includes:

- 3 (A) sum input means for receiving said digital sum signal;
4 (B) sum output means for providing a conditioned sum signal;
5 (C) a sum signal transmission path coupled to said sum input means
6 and to said sum output means, and including means for generating said
7 conditioned sum signal from said digital sum signal.

1 31. A digital BTSC encoder according to claim 30, wherein said sum channel
2 processing means includes sum channel low pass filter means for receiving a sum
3 channel low pass filter input signal from said sum signal transmission path and
4 including means for low pass filtering said sum channel low pass filter input
5 signal and thereby generating a sum channel low pass filter output signal, and
6 including means for applying said sum channel low pass filter output signal to
7 said sum signal transmission path.

1 32. A digital BTSC encoder according to claim 31, wherein the filtering
2 provided by a cascade of said first and second low pass filter means is
3 substantially similar to the filtering provided by said sum channel low pass
4 filtering means.

1 33. A digital BTSC encoder according to claim 32, wherein the filtering
2 provided by said sum channel low pass filtering means is characterized by a null
3 at 15,734 Hz.

4 34. A digital BTSC encoder according to claim 33, wherein the filtering
5 provided by said sum channel low pass filtering means is characterized by a pass
6 band between zero and 15 kHz

1 35. A digital BTSC encoder according to claim 34, wherein the filtering





2 provided by said sum channel low pass filtering means is characterized by a
3 cutoff above 15 kHz.

1 36. A digital BTSC encoder according to claim 31, wherein said difference
2 channel processing means includes first fixed preemphasis filter means for
3 receiving a first preemphasis input signal from said difference signal transmission
4 path and including means for filtering said first preemphasis input signal and
5 thereby generating a first preemphasis output signal, and including means for
6 applying said first preemphasis output signal to said difference signal transmission
7 path.

1 37. A digital BTSC encoder according to claim 36, wherein said difference
2 channel processing means includes second fixed preemphasis filter means for
3 receiving a second preemphasis input signal from said difference signal
4 transmission path and including means for filtering said second preemphasis input
5 signal and thereby generating a second preemphasis output signal, and including
6 means for applying said second preemphasis output signal to said difference signal
7 transmission path.

1 38. A digital BTSC encoder according to claim 37, wherein said sum channel
2 processing means includes 75 μ s preemphasis filter means for receiving a 75 μ s
3 preemphasis input signal from said sum signal transmission path and including
4 means for filtering said 75 μ s preemphasis input signal and thereby generating a
5 75 μ s preemphasis output signal, and including means for applying said 75 μ s
6 preemphasis output signal to said sum signal transmission path.

1 39. A digital BTSC encoder according to claim 38, wherein said sum channel
2 processing means includes static equalization filter means for receiving a static
3 equalization input signal from said sum signal transmission path and including
4 means for filtering said static equalization input signal and thereby generating a
5 static equalization output signal, and including means for applying said static
6 equalization output signal to said sum signal transmission path.





7 40. A digital BTSC encoder according to claim 39, wherein the filtering
8 provided by said second preemphasis filter means is characterized by a second
9 preemphasis phase response, the filtering provided by said 75 μ s preemphasis
10 filter means being characterized by a 75 μ s preemphasis phase response, a
11 difference between said second preemphasis phase response and a first reference
12 phase response being substantially similar to a difference between said 75 μ s
13 preemphasis phase response and a second reference phase response.

1 41. A digital BTSC encoder according to claim 40, wherein the filtering
2 provided by said first preemphasis filter means is characterized by a first
3 preemphasis phase response, the filtering provided by said static equalization
4 filter means being characterized by a static equalization phase response, said first
5 preemphasis phase response being substantially similar to said static equalization
6 phase response.

1 42. A digital BTSC encoder according to claim 41, wherein said first
2 preemphasis input signal comprises said first low pass filter output signal.

1 43. A digital BTSC encoder according to claim 42, wherein said wideband
2 compression input signal comprises said first preemphasis output signal.

1 44. A digital BTSC encoder according to claim 43, wherein said second
2 preemphasis input signal comprises said spectral compression output signal.

3 45. A digital BTSC encoder according to claim 41, wherein said difference
4 channel processing means includes difference dynamic equalization filter means
5 for receiving a difference dynamic equalization input signal from said difference
6 signal transmission path and including means for filtering said difference dynamic
7 equalization input signal and thereby generating a difference dynamic equalization
8 output signal, and including means for applying said difference dynamic
9 equalization output signal to said difference signal transmission path.



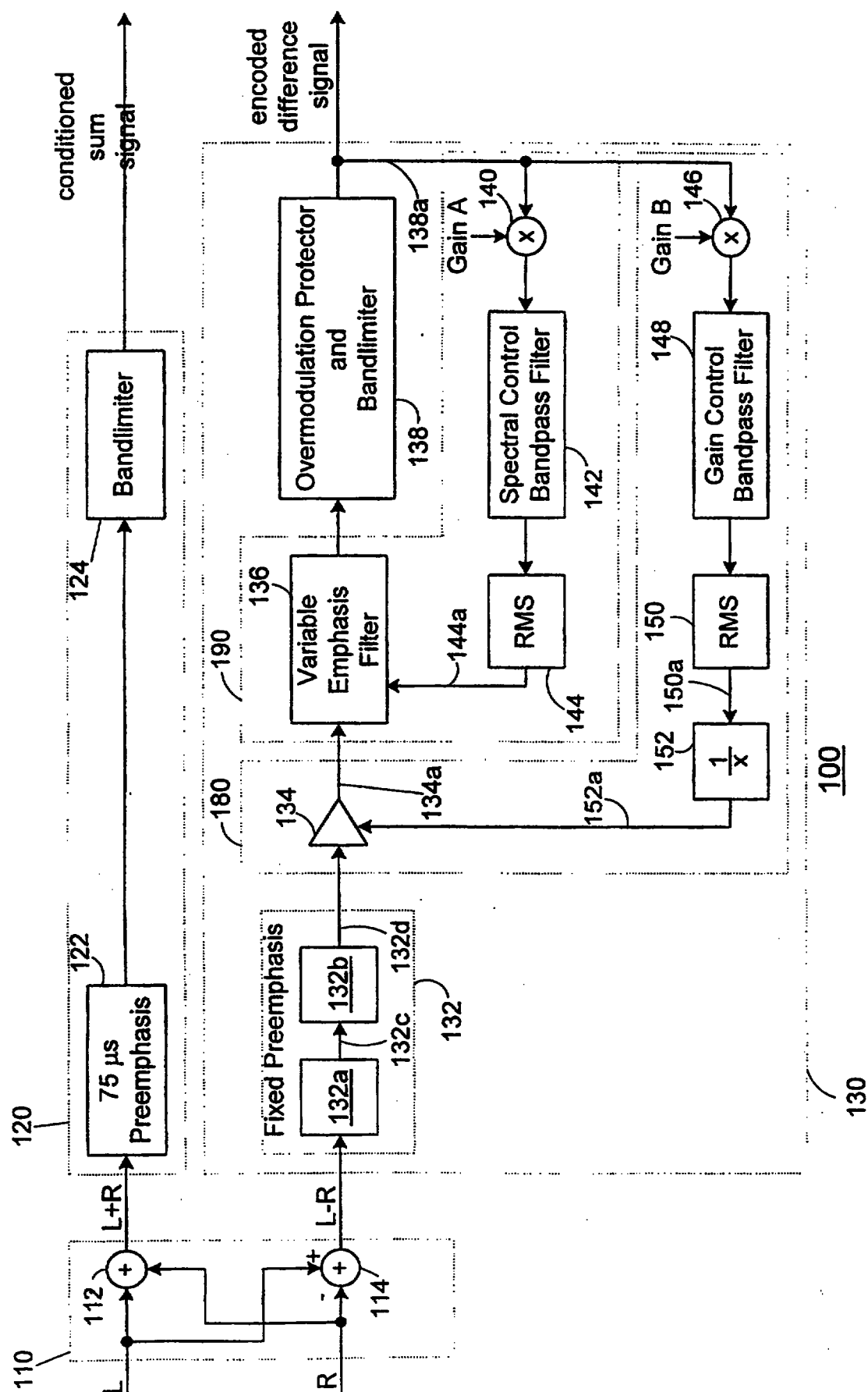


1 46. A digital BTSC encoder according to claim 45, wherein said sum channel
2 processing means includes sum dynamic equalization filter means for receiving a
3 sum dynamic equalization input signal from said sum signal transmission path and
4 including means for filtering said sum dynamic equalization input signal and
5 thereby generating a sum dynamic equalization output signal, and including means
6 for applying said sum dynamic equalization output signal to said sum signal
7 transmission path.

1 47. A digital BTSC encoder according to claim 46, wherein the filtering
2 provided by said difference dynamic equalization filter means is characterized by
3 a difference dynamic equalization phase response, the filtering provided by said
4 variable emphasis filter being characterized by a variable emphasis phase
5 response, the filtering provided by said sum dynamic equalization filter means
6 being characterized by a sum dynamic equalization phase response, said
7 difference dynamic equalization, variable emphasis, and sum dynamic
8 equalization phase responses all varying according to a function of said second
9 control signal.

1 48. A digital BTSC encoder according to claim 30, further including
2 composite modulator means for receiving said encoded difference signal and said
3 conditioned sum signal and for generating therefrom a composite modulated
4 signal.





Prior Art
Fig. 1

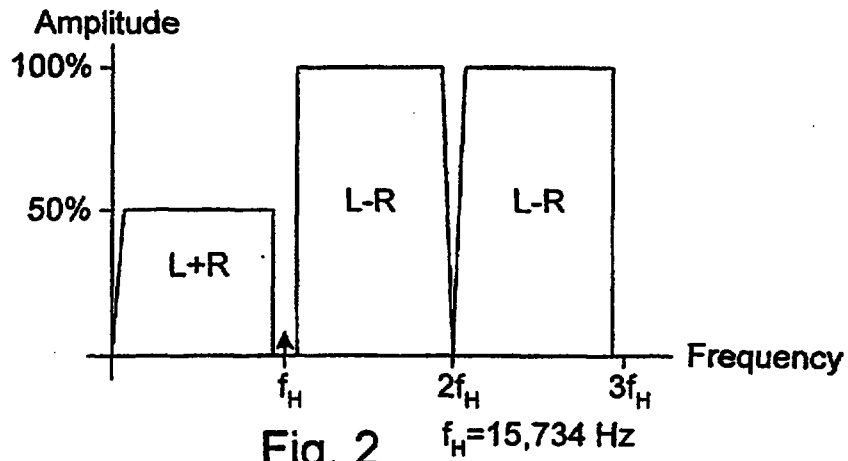


Fig. 2

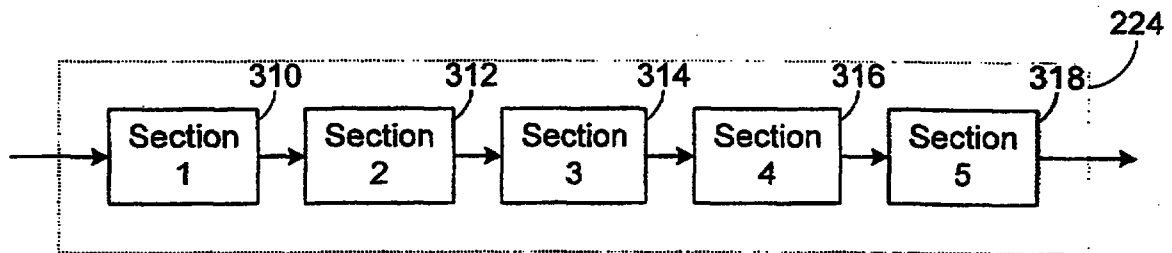


Fig. 4A

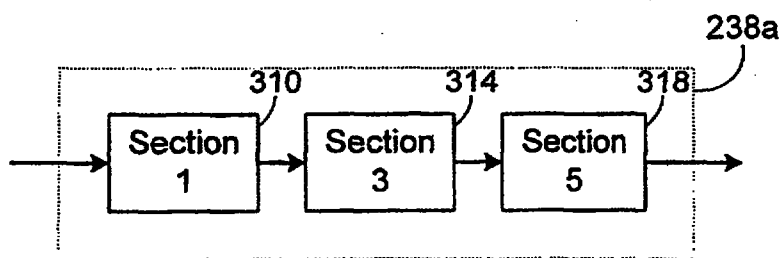


Fig. 4B

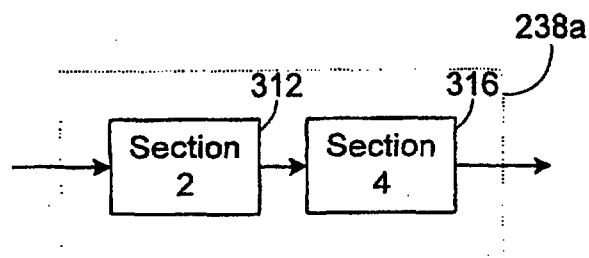
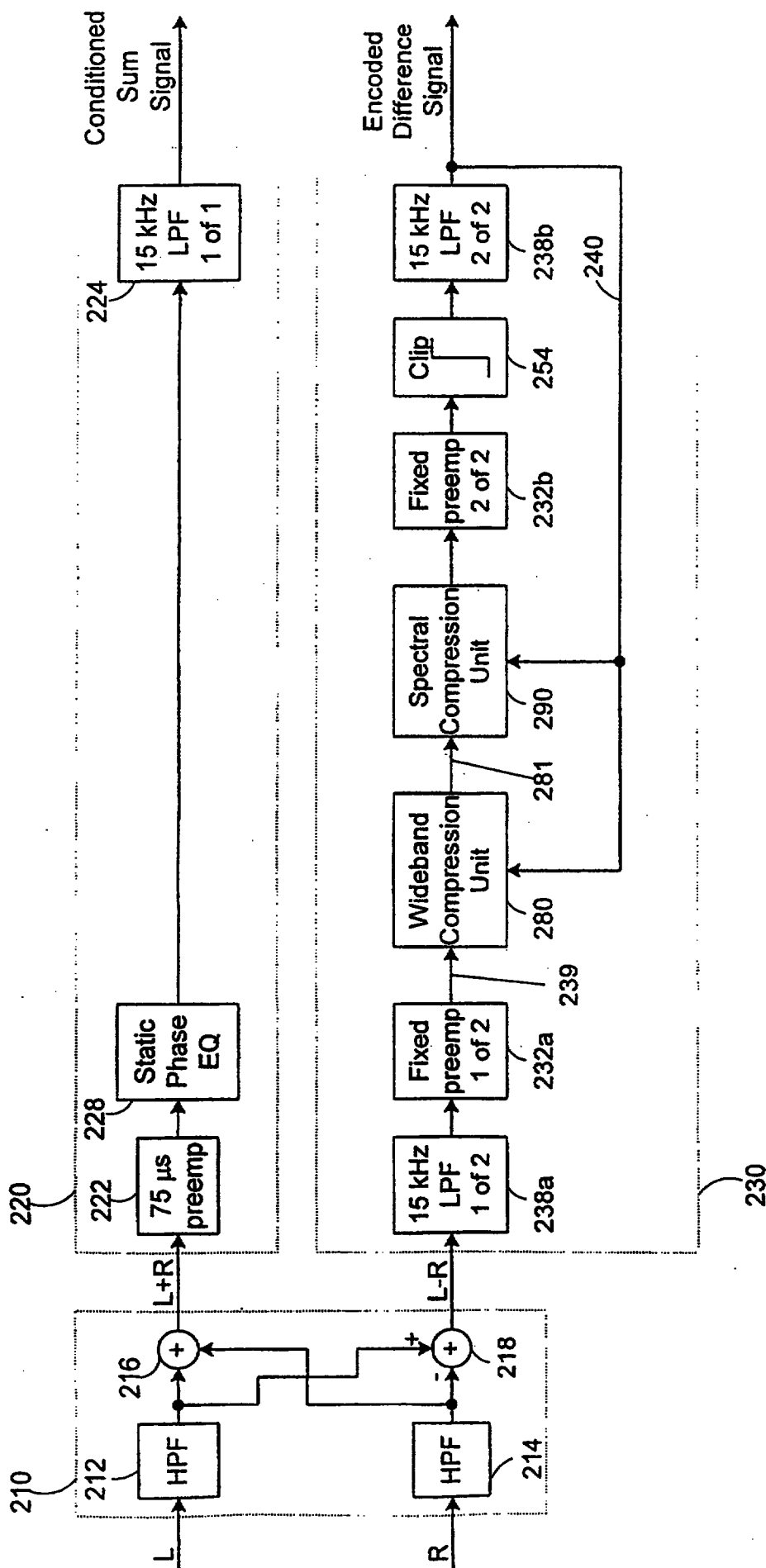


Fig. 4C



200

Fig. 3

4/10

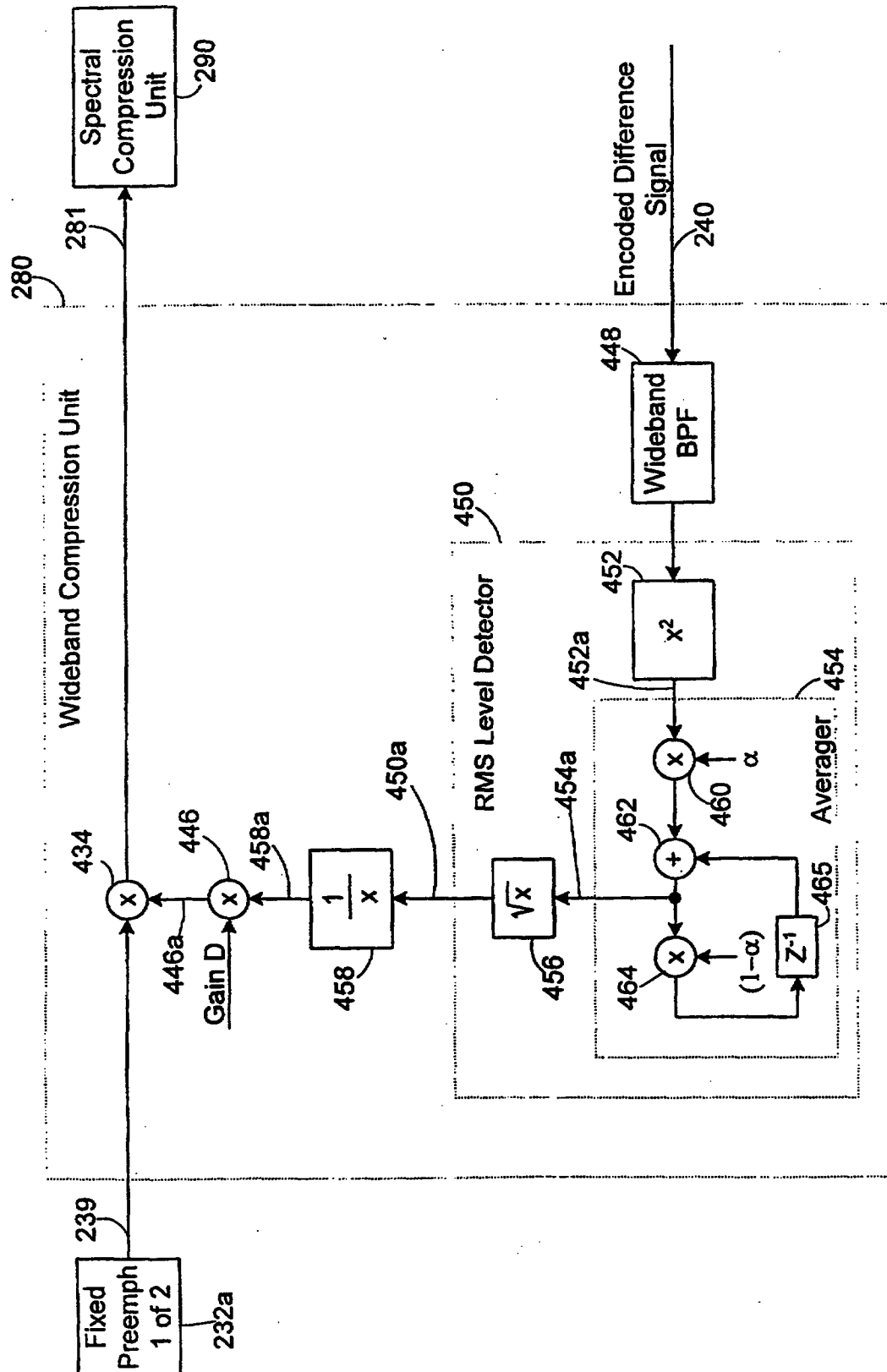


Fig. 5

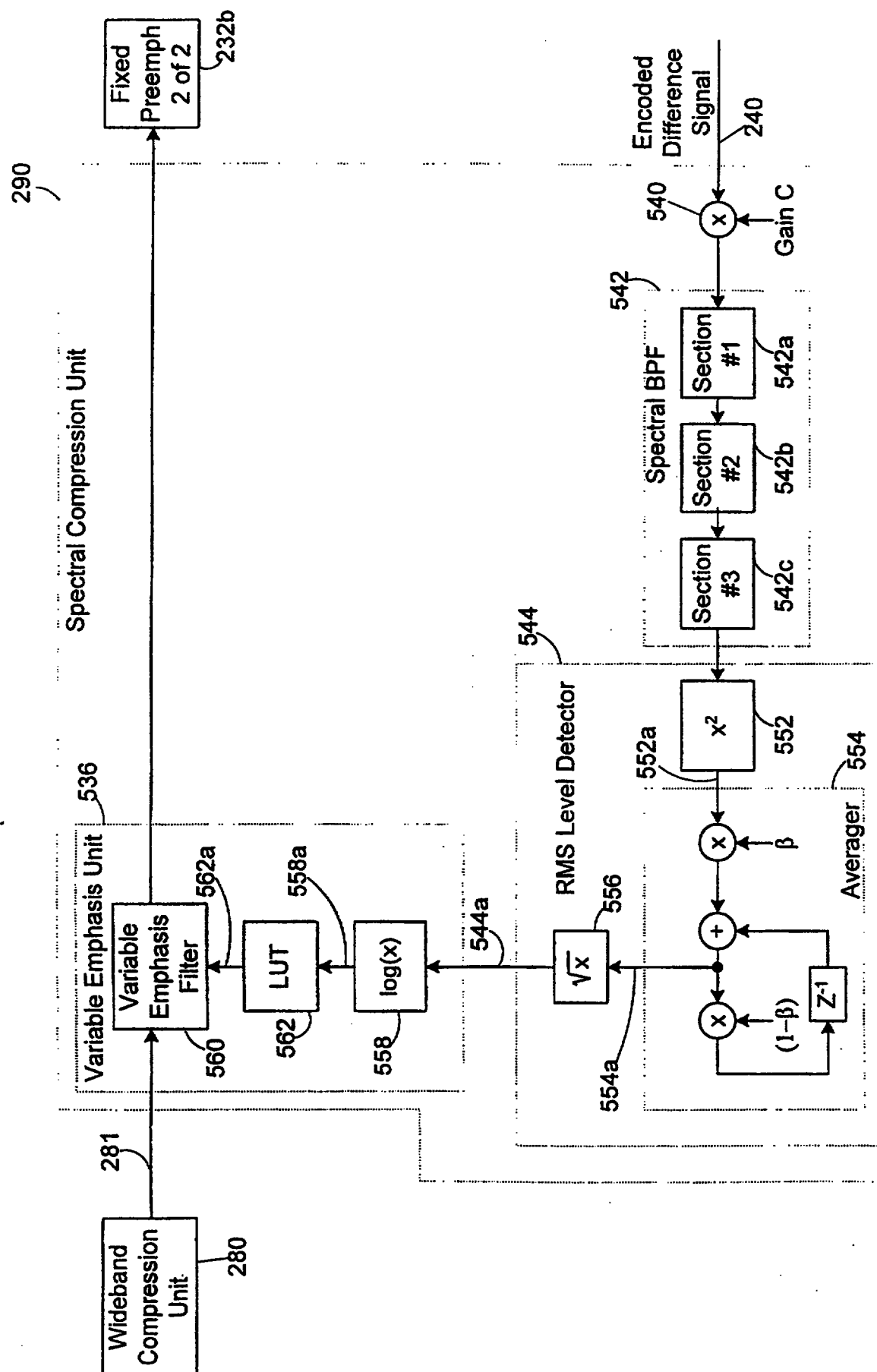
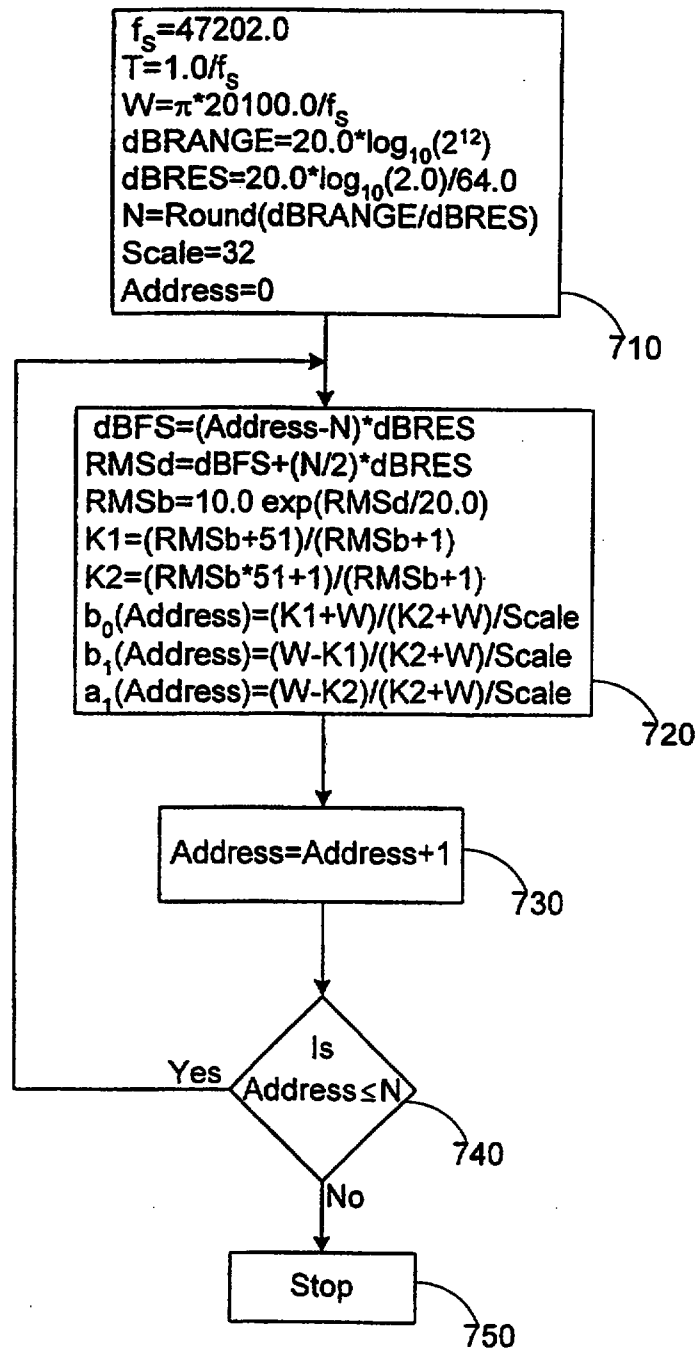


Fig. 6

6/10



700

Fig. 7

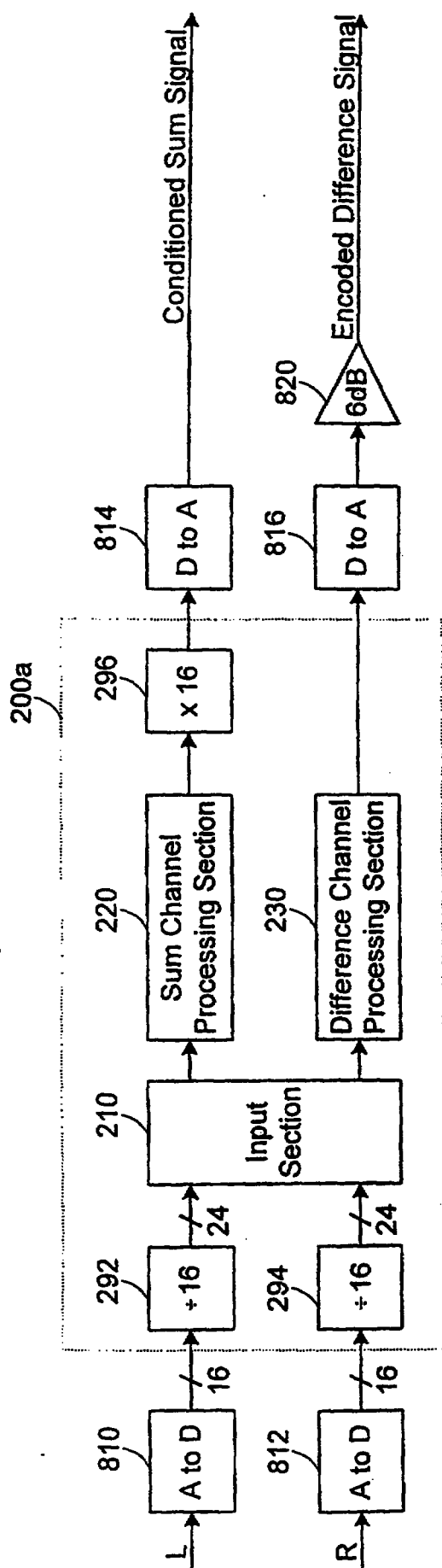


Fig. 8A

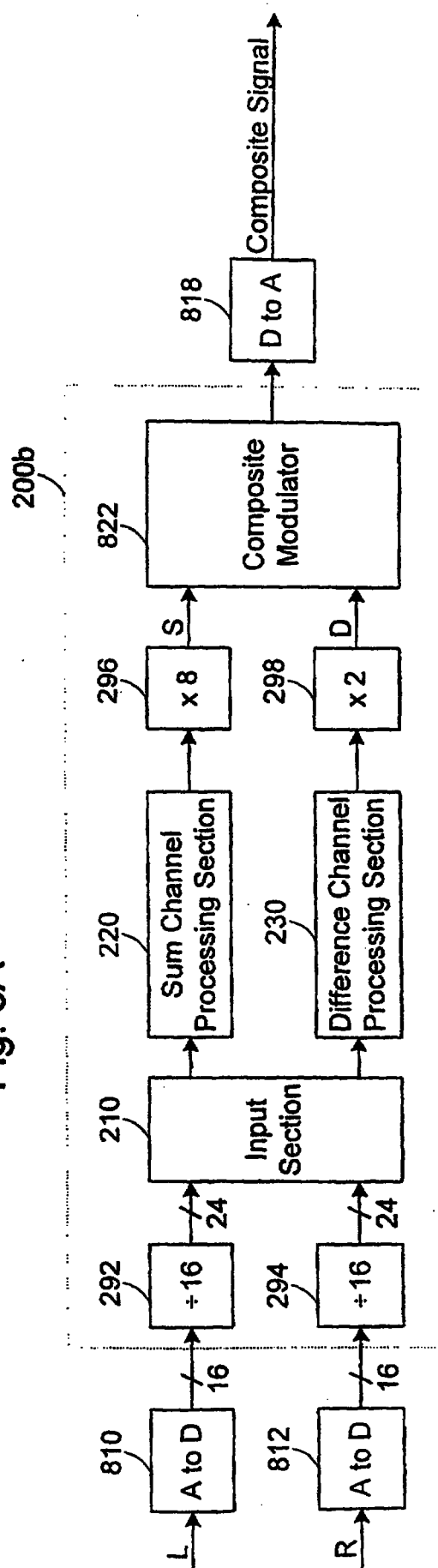


Fig. 8B

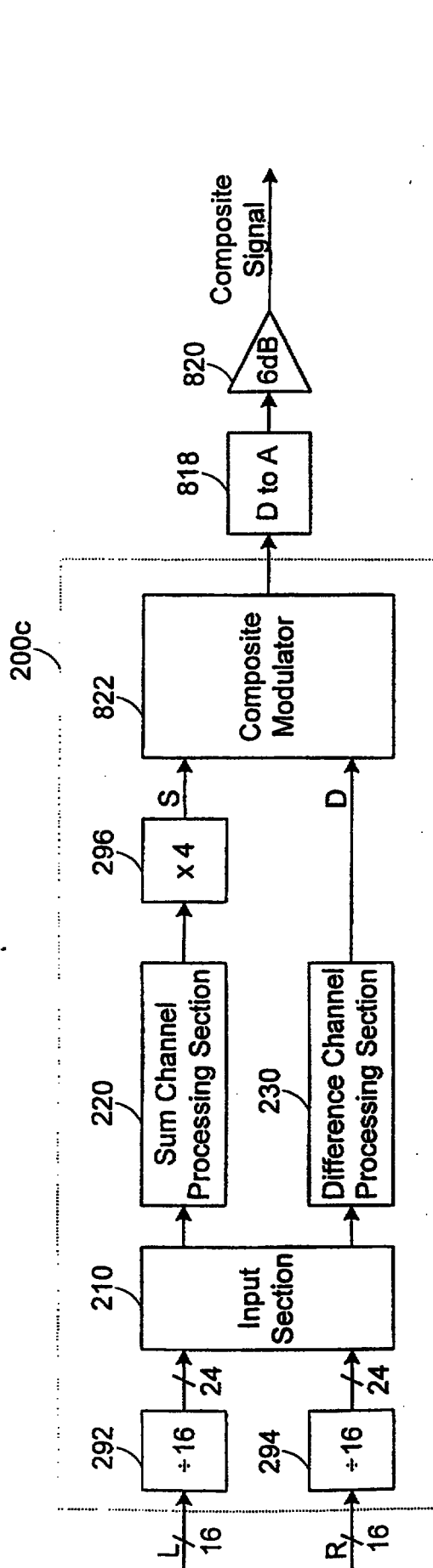


Fig. 8C

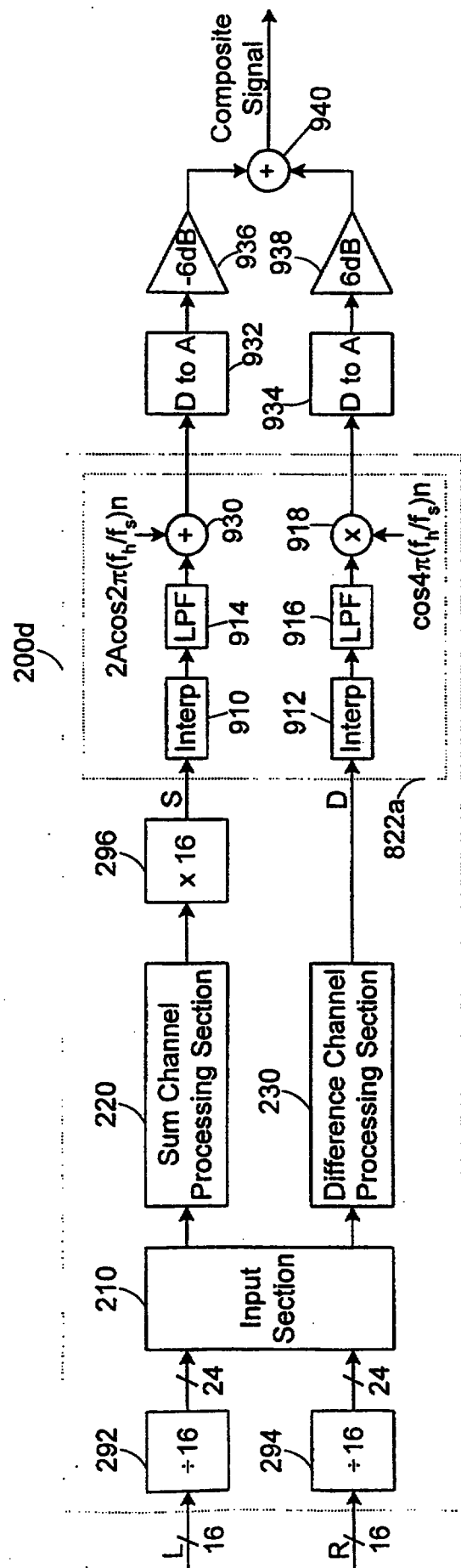


Fig. 8D

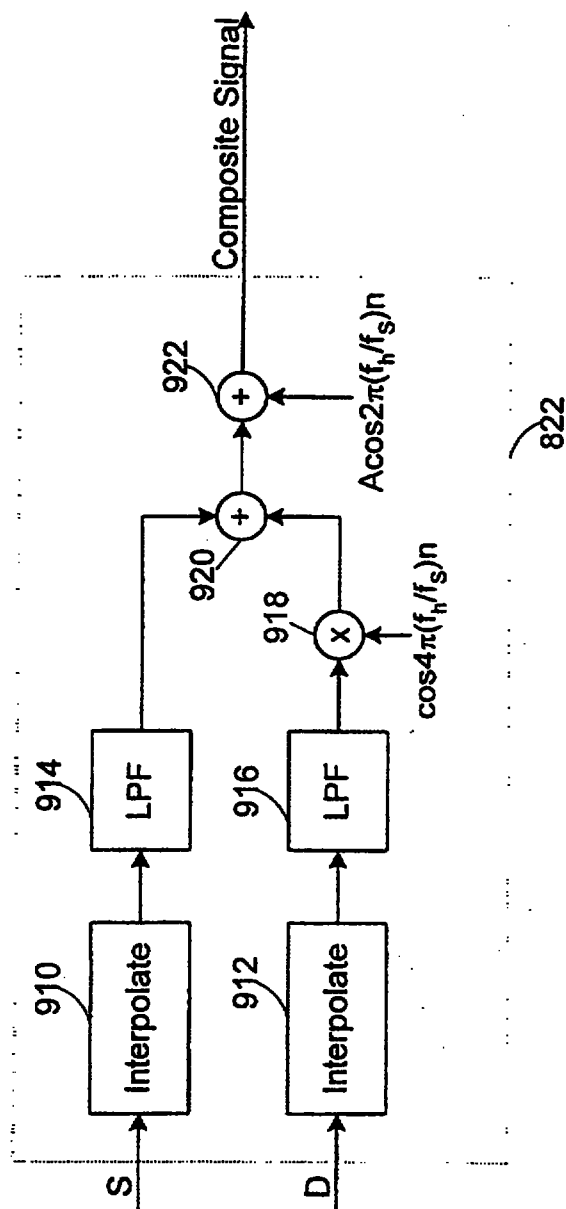


Fig. 9

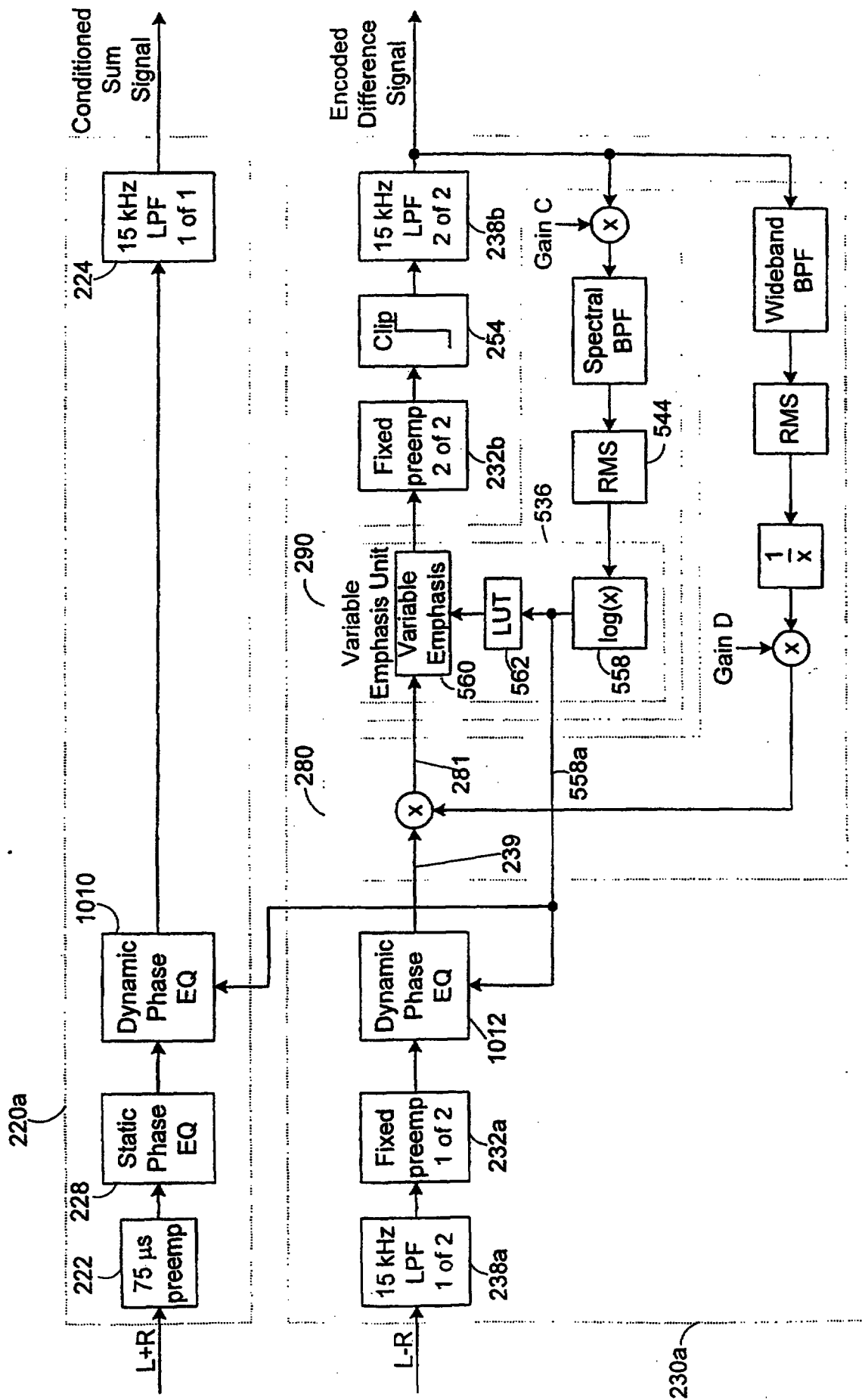
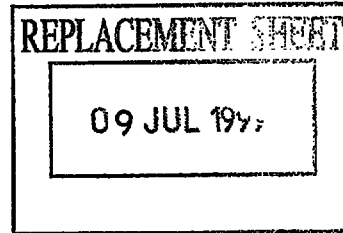


Fig. 10



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CLAIMS

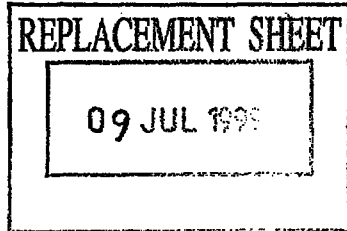
1 1. An adaptive signal weighting assembly including a signal path for transmitting
2 an information signal of a predetermined bandwidth through the assembly, the assembly
3 further comprising: (A) filter means disposed in the signal path for varying the gain impressed
4 on the portion of the information signal within a first select spectral region within the
5 predetermined bandwidth by a first variable gain factor, the first variable gain factor varying
6 in response to and as a function of a first control signal; (B) first control signal generating
7 means for generating the first control signal only in response to and in accordance with the
8 signal energy of the information signal within a second select spectral region including at least
9 a part of the first select spectral region; (C) gain control means disposed in the signal path and
10 coupled to the filter means for varying the signal gain impressed on the information signal
11 substantially throughout the predetermined bandwidth by a second variable gain factor, the
12 second variable gain factor varying in response to and as a function of a second control signal;
13 and (D) second control signal generating means for generating the second control signal in
14 response to and as a function of the signal energy of the information signal substantially
15 within a third select spectral region within the predetermined bandwidth; wherein the
16 assembly is characterized by:

17 the information signal being a digital information signal;

18 the filter means being a digital filter disposed in the signal path for varying the gain
19 impressed on the portion of the digital information signal within the first select spectral region
20 within the predetermined bandwidth by the first variable gain factor;

21 the gain control means being a digital gain control means disposed in the signal path
22 and coupled to the digital filter means for varying the signal gain impressed on the digital
23 information signal substantially throughout the predetermined bandwidth;

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24 the second control generating means digitally generating the second control signal in
25 response to and as a function of the signal energy of the digital information signal
26 substantially within the third select spectral region within the predetermined bandwidth.

27 wherein the digital filter means, first control signal generating means, digital gain
28 control means, and the second control signal generating means each operate in substantially
29 real time so as to preserve the signal content of information signal.

1 2. An adaptive signal weighting assembly according to claim 1, wherein the
2 assembly encodes the information signal, and further including: (A) input means for receiving
3 the information signal; and (B) output means coupled to the input means through the signal
4 path for providing the information signal as encoded by the assembly; the system further
5 characterized by:

6 the filter means including variable coefficient digital filter means for filtering the
7 information signal, the filtering being characterized by a variable coefficient transfer function
8 and the filtering varying the gain impressed on the portion of the information signal within the
9 first select spectral region within the predetermined bandwidth by the first variable gain
10 factor, the variable coefficients of the variable coefficient transfer function and the first
11 variable gain factor varying in response to and as a function of the first control signal.

1 3. An adaptive signal weighting assembly according to claim 1, further
2 characterized by

3 the assembly being adapted to encode in real time two information signals comprising
4 digital left and right channel audio signals so that the encoded left and right channel audio
5 signals can be subsequently decoded so as to reproduce the digital left and right channel audio
6 signals with little or no distortion of the signal content of the digital left and right channel
7 audio signals.

1 4. An adaptive signal weighting assembly according to claim 3, further characterized
2 by:

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3 (A) left high pass filter means for receiving the digital left channel audio signal and
4 for digitally high pass filtering the digital left channel audio signal and thereby generating a
5 digital left filtered signal;

6 (B) right high pass filter means for receiving the digital right channel audio signal
7 and for digitally high pass filtering the digital right channel audio signal and thereby
8 generating a digital right filtered signal;

9 (C) matrix means for receiving the digital left and digital right filtered signals, and
10 including means for summing the digital left and digital right filtered signals and thereby
11 generating a digital sum signal, and including means for subtracting one of the digital left and
12 digital right filtered signals from the other of the digital left and digital right filtered signals
13 and thereby generating a digital difference signal;

14 (D) difference channel processing means for digitally processing the digital
15 difference signal; and

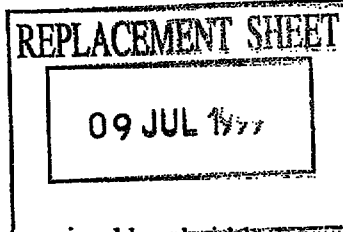
16 (E) sum channel processing means for digitally processing the digital sum signal;
17 wherein the left and right high pass filter means, the matrix means, and the difference
18 channel processing means and the sum channel processing means each operate in substantially
19 real time so as to preserve the signal content of the digital left and right channel audio signals.

1 5. An adaptive signal weighting assembly according to claim 4, further
2 characterized wherein the left and right high pass filter means are each characterized by a
3 cutoff frequency that is less than or equal to 50 Hz.

1 6. An adaptive signal weighting assembly according to claim 5, further
2 characterized wherein the left and right high pass filter means are each characterized by a
3 passband and a substantially flat response in the passband.

1 7. An adaptive digital signal weighting system for use with a digitally represented
2 difference signal representing the difference between two stereophonic audio signals, the
3 system including a signal path for transmitting an electrical information signal containing

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information relating to the difference signal of a predetermined bandwidth through said system at a predetermined sample rate so as to preserve the signal content of said information signal, said system further comprising:

a digital filter arrangement disposed in said signal path and constructed and arranged so as to vary the gain impressed on the portion of said information signal within a first select spectral region within said predetermined bandwidth by a first variable gain factor, said first variable gain factor varying in response to and as a function of a first control signal;

a first control signal generator constructed and arranged so as to digitally generate said first control signal only in response to and in accordance with the signal energy of said digitally represented difference signal within a second select spectral region including at least a part of said first select spectral region;

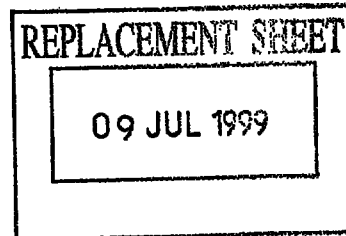
a digital gain controller, disposed in said signal path and coupled to said digital filter arrangement, and constructed and arranged so as to vary the signal gain impressed on said information signal substantially throughout said predetermined bandwidth by a second variable gain factor, said second variable gain factor varying in response to and as a function of a second control signal; and

a second control signal generator constructed and arranged so as to digitally generate said second control signal in response to and as a function of the signal energy of said digitally represented difference signal substantially within a third select spectral region within said predetermined bandwidth;

wherein the digital filter arrangement, first control signal generator, digital gain controller, and the second control signal generator each operate at a predetermined sample rate so as to preserve the signal content of the information signal, and the sampling rate is chosen so as to be equal to an integer multiple of the frequency of a pilot tone that can be added to the difference signal to identify the encoded signal to a receiver.

8. A digital system for encoding an electrical information signal of a predetermined bandwidth at a predetermined sample rate so as to preserve the signal content of said information and so that said information signal can be recorded on or transmitted

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4 through a channel with a frequency dependent dynamic range as an encoded difference signal
5 representing the difference between two stereophonic audio signals, the channel having a
6 narrower dynamically-limited portion in a first spectral region than in at least one other
7 spectral region of said predetermined bandwidth, said system comprising:

8 an input constructed and arranged so as to receive said information signal;

9 a signal transmission path, coupled to said input and constructed and arranged so as to
10 transmit said information signal received at said input;

11 an output coupled to said input through said signal transmission path for providing
12 said information signal as encoded by said system;

13 a digital gain controller coupled to said signal path for varying the signal gain
14 impressed on said information signal substantially throughout said predetermined bandwidth,
15 said signal gain varying in response to and as a function of a first control signal;

16 a digital filter arrangement, coupled to said signal path and said digital gain controller,
17 and constructed and arranged so as to impress a second variable gain on the portion of said
18 information signal substantially within said first spectral region so as to preemphasize said
19 portion with respect to the remaining portions of said information signal, said second variable
20 gain varying in response to and as a function of a second control signal;

21 a first control signal generator constructed and arranged so as to digitally generate said
22 first control signal in response to and as a function of the signal energy of said information
23 signal substantially within a second spectral region of said predetermined bandwidth; and

24 a second control signal generator constructed and arranged so as to digitally generate
25 said second control signal only in response to and in accordance with the signal energy of said
26 information signal within a third spectral region of said predetermined bandwidth including at
27 least a part of said first spectral region;

28 wherein the digital gain controller, said digital filter arrangement, said first control
29 signal generator and said second control signal generator each operate at a predetermined
30 sample rate so as to preserve the signal content of said information signal, and the sampling
31 rate is chosen so as to be equal to an integer multiple of the frequency of a pilot tone that can
32 be added to the encoded signal to identify the encoded signal to a receiver.

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9. An adaptive digital signal weighting system for use with a digitally represented difference signal representing the difference between two stereophonic audio signals, the system comprising:

a signal path for transmitting an electrical information signal of a predetermined bandwidth through said system;

a variable coefficient digital filter arrangement (a) constructed and arranged so as to filter said information signal, the filter arrangement being characterized by a variable coefficient transfer function, and (b) constructed and arranged so as to vary the gain impressed on the portion of said information signal within a first select spectral region within said predetermined bandwidth by a first variable gain factor, the variable coefficients of said variable coefficient transfer function and said first variable gain factor varying in response to and as a function of a first control signal;

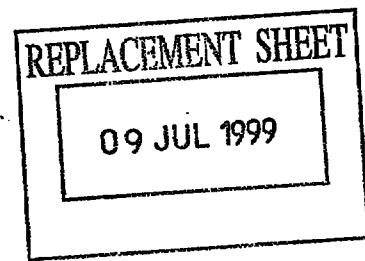
a first control signal generator constructed and arranged so as to digitally generate said first control signal only in response to and in accordance with the signal energy of said information signal within a second select spectral region including at least a part of said first select spectral region;

a digital gain controller disposed in said signal path and coupled to said variable coefficient digital filter arrangement, and constructed and arranged so as to vary the signal gain impressed on said information signal substantially throughout said predetermined bandwidth by a second variable gain factor, said second variable gain factor varying in response to and as a function of a second control signal; and

a second control signal generator constructed and arranged so as to digitally generate said second control signal in response to and as a function of the signal energy of said information signal substantially within a third select spectral region within said predetermined bandwidth;

wherein the digital filter arrangement, first control signal generator, digital gain controller and second control signal generator each operate at a predetermined sample rate so as to preserve the signal content of said information signal, and the sampling rate is chosen so as to be equal to an integer multiple of the frequency of a pilot tone that can be added to the

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30 difference signal to identify the difference signal to a receiver.

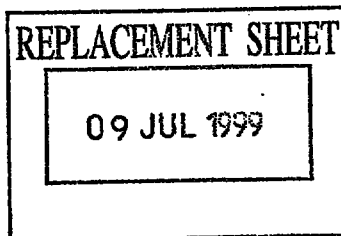
1 10. An adaptive signal weighting assembly according to claim 9, further
2 characterized wherein the sum channel processing means and the difference channel
3 processing means are both implemented on a single integrated circuit.

1 11. An adaptive signal weighting assembly according to claim 3, further
2 characterized by
3 the digital left and right channel audio signals are digitally sampled signals sampled
4 with a sampling frequency substantially equal to N times 15,734 Hz, N being an integer
5 greater than or equal to three.

1 12. An adaptive signal weighting assembly according to claim 9, further
2 characterized by:
3 the difference channel processing means including:
4 (A) difference input means for receiving the digital difference signal;
5 (B) difference output means for providing an encoded difference signal;
6 (C) a difference signal transmission path coupled to the difference input means and
7 to the difference output means, and including means for generating the encoded difference
8 signal from the digital difference signal.

1 13. An adaptive signal weighting assembly according to claim 12, further
2 characterized wherein the difference channel processing means further includes spectral
3 compression means for receiving a spectral compression input signal from the difference
4 signal transmission path and for compressing the spectral compression input signal according
5 to a function of an energy level of the encoded difference signal and thereby generating a
6 spectral compression output signal, and including means for applying the spectral
7 compression output signal to the difference signal transmission path.

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1 14. An adaptive signal weighting assembly according to claim 13, further
2 characterized wherein the spectral compression means includes means for measuring a first
3 energy level of the encoded difference signal in a first select spectral portion, and for
4 generating a first control signal representative of the first energy level.

1 15. An adaptive signal weighting assembly according to claim 14, further
2 characterized wherein the spectral compression means further includes a variable emphasis
3 filter for receiving and digitally filtering the spectral compression input signal and thereby
4 generating the spectral compression output signal, the filtering provided by the variable
5 emphasis filter being characterized by a transfer function including a plurality of coefficients,
6 the spectral compression means further including means for selecting the plurality of
7 coefficients according to a function of the first control signal.

1 16. An adaptive signal weighting assembly according to claim 15, further
2 characterized wherein the means for selecting the plurality of coefficients includes a memory
3 look up table.

1 17. An adaptive signal weighting assembly according to claim 16, further
2 characterized wherein the means for selecting the plurality of coefficients includes (a)
3 logarithmic generator means for receiving and logarithmically compressing the first control
4 signal and thereby generating a logarithmically compressed signal and (b) means for applying
5 the logarithmically compressed signal to the memory look up table.

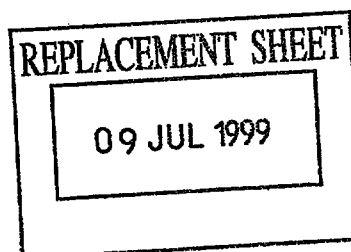
1 18. An adaptive signal weighting assembly according to claim 15, further
2 characterized wherein the means for selecting the plurality of coefficients includes means for
3 calculating the plurality of coefficients as a function of the first control signal.

1 19. An adaptive signal weighting assembly according to claim 15, further
2 characterized wherein the spectral compression means includes spectral bandpass filter means

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3 for receiving and filtering a signal representative of the encoded difference signal and thereby
4 generating a spectral signal, the filtering provided by the spectral bandpass filter means being
5 characterized by a passband in the first select spectral portion.

1 20. An adaptive signal weighting assembly according to claim 19, further
2 characterized wherein the spectral compression means includes first RMS level detector
3 means for receiving the spectral signal and for generating therefrom the first control signal,
4 the first control signal being representative of an RMS value of the spectral signal.

1 21. An adaptive signal weighting assembly according to claim 19, further
2 characterized wherein the spectral compression means includes amplifying means for
3 receiving and amplifying the encoded difference signal and thereby generating the signal
4 representative of the encoded difference signal.

1 22. An adaptive signal weighting assembly according to claim 15, further
2 characterized wherein the difference channel processing means includes wideband
3 compression means for receiving a wideband compression input signal from the difference
4 signal transmission path and for compressing the wideband compression input signal
5 according to a function of an energy level of the encoded difference signal and thereby
6 generating a wideband compression output signal, and including means for applying the
7 wideband compression output signal to the difference signal transmission path.

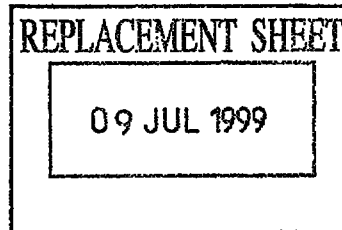
1 23. An adaptive signal weighting assembly according to claim 22, further
2 characterized wherein the spectral compression input signal comprises the wideband
3 compression output signal.

1 24. An adaptive signal weighting assembly according to claim 22, further
2 characterized wherein the wideband compression means includes means for measuring a
3 second energy level of the encoded difference signal in a second select spectral portion, and

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4 for generating a second control signal representative of the second energy level.

1 25. An adaptive signal weighting assembly according to claim 24, further
2 characterized wherein the wideband compression means includes amplifier means for
3 receiving and amplifying the wideband compression input signal using a gain controlled by
4 the second control signal and thereby generating the wideband compression output signal.

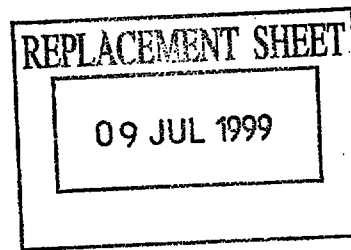
1 26. An adaptive signal weighting assembly according to claim 25, further
2 characterized wherein the wideband compression means further includes wideband bandpass
3 filter means for receiving and filtering a signal representative of the encoded difference signal
4 and thereby generating a wideband signal, the filtering provided by the wideband bandpass
5 filter means being characterized by a passband in the second select spectral portion.

1 27. An adaptive signal weighting assembly according to claim 26, further
2 characterized wherein the wideband compression means further includes second RMS level
3 detector means for receiving the wideband signal and for generating therefrom the second
4 control signal, the second control signal being representative of an RMS value of the
5 wideband signal.

1 28. An adaptive signal weighting assembly according to claim 22, further
2 characterized wherein the difference channel processing means further includes first low pass
3 filter means for receiving a first low pass filter input signal from the difference signal
4 transmission path and including means for low pass filtering the first low pass filter input
5 signal and thereby generating a first low pass filter output signal, and including means for
6 applying the first low pass filter output signal to the difference signal transmission path.

1 29. An adaptive signal weighting assembly according to claim 28, further
2 characterized wherein the difference channel processing means further includes second low
3 pass filter means for receiving a second low pass filter input signal from the difference signal





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transmission path and including means for low pass filtering the second low pass filter input signal and thereby generating a second low pass filter output signal, and including means for applying the second low pass filter output signal to the difference signal transmission path.

30. An adaptive signal weighting assembly according to claim 29, further characterized wherein the sum channel processing means includes:

- (A) sum input means for receiving the digital sum signal;
- (B) sum output means for providing a conditioned sum signal;
- (C) a sum signal transmission path coupled to the sum input means and to the sum output means, and including means for generating the conditioned sum signal from the digital sum signal.

31. An adaptive signal weighting assembly according to claim 30, further characterized wherein the sum channel processing means further includes (a) sum channel low pass filter means for receiving a sum channel low pass filter input signal from the sum signal transmission path and (b) means for low pass filtering the sum channel low pass filter input signal and thereby generating a sum channel low pass filter output signal, and including means for applying the sum channel low pass filter output signal to the sum signal transmission path.

32. An adaptive signal weighting assembly according to claim 31, further characterized wherein the filtering provided by a cascade of the first and second low pass filter means is substantially similar to the filtering provided by the sum channel low pass filtering means.

33. An adaptive signal weighting assembly according to claim 32, further characterized wherein the filtering provided by the sum channel low pass filtering means is characterized by a null at 15,734 Hz.

34. An adaptive signal weighting assembly according to claim 33, further

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2 characterized wherein the filtering provided by the sum channel low pass filtering means is
3 characterized by a pass band between zero and 15 kHz

1 35. An adaptive signal weighting assembly according to claim 34, further
2 characterized wherein the filtering provided by the sum channel low pass filtering means is
3 characterized by a cutoff above 15 kHz.

1 36. An adaptive signal weighting assembly according to claim 31, further
2 characterized wherein the difference channel processing means further includes (a) first fixed
3 preemphasis filter means for receiving a first preemphasis input signal from the difference
4 signal transmission path, (b) means for filtering the first preemphasis input signal so as to
5 generate a first preemphasis output signal, and (c) means for applying the first preemphasis
6 output signal to the difference signal transmission path.

1 37. An adaptive signal weighting assembly according to claim 36, further
2 characterized wherein the difference channel processing means further includes (a) second
3 fixed preemphasis filter means for receiving a second preemphasis input signal from the
4 difference signal transmission path, (b) means for filtering the second preemphasis input
5 signal and thereby generating a second preemphasis output signal, and (c) means for applying
6 the second preemphasis output signal to the difference signal transmission path.

1 38. An adaptive signal weighting assembly according to claim 37, further
2 characterized wherein the sum channel processing means further includes (a) 75 μ s
3 preemphasis filter means for receiving a 75 μ s preemphasis input signal from the sum signal
4 transmission path, (b) means for filtering the 75 μ s preemphasis input signal so as to generate
5 a 75 μ s preemphasis output signal, and (c) means for applying the 75 μ s preemphasis output
6 signal to the sum signal transmission path.

1 39. An adaptive signal weighting assembly according to claim 38, further

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2 characterized wherein the sum channel processing means further includes (a) static
3 equalization filter means for receiving a static equalization input signal from the sum signal
4 transmission path, (b) means for filtering the static equalization input signal so as to a static
5 equalization output signal, and (c) means for applying the static equalization output signal to
6 the sum signal transmission path.

1 40. An adaptive signal weighting assembly according to claim 39, further
2 characterized wherein the filtering provided by the second preemphasis filter means is
3 characterized by a second preemphasis phase response, the filtering provided by the 75 μ s
4 preemphasis filter means is characterized by a 75 μ s preemphasis phase response, and a
5 difference between the second preemphasis phase response and a first reference phase
6 response is substantially similar to a difference between the 75 μ s preemphasis phase response
7 and a second reference phase response.

1 41. An adaptive signal weighting assembly according to claim 40, further
2 characterized wherein the filtering provided by the first preemphasis filter means is
3 characterized by a first preemphasis phase response, the filtering provided by the static
4 equalization filter means is characterized by a static equalization phase response, and the first
5 preemphasis phase response is substantially similar to the static equalization phase response.

1 42. An adaptive signal weighting assembly according to claim 41, further
2 characterized wherein the first preemphasis input signal comprises the first low pass filter
3 output signal.

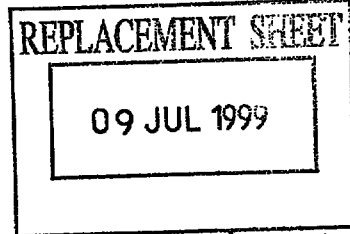
1 43. An adaptive signal weighting assembly according to claim 42, further
2 characterized wherein the wideband compression input signal comprises the first preemphasis
3 output signal.

1 44. An adaptive signal weighting assembly according to claim 43, further

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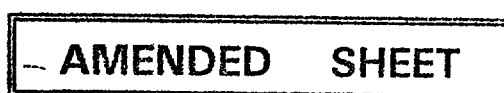
2 characterized wherein the second preemphasis input signal comprises the spectral
3 compression output signal.

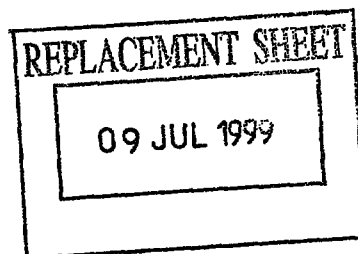
1 45. An adaptive signal weighting assembly according to claim 41, further
2 characterized wherein the difference channel processing means further includes (a) difference
3 dynamic equalization filter means for receiving a difference dynamic equalization input signal
4 from the difference signal transmission path, (b) means for filtering the difference dynamic
5 equalization input signal so as to generate a difference dynamic equalization output signal,
6 and (c) means for applying the difference dynamic equalization output signal to the difference
7 signal transmission path.

1 46. An adaptive signal weighting assembly according to claim 45, further
2 characterized wherein the sum channel processing means further includes (a) sum dynamic
3 equalization filter means for receiving a sum dynamic equalization input signal from the sum
4 signal transmission path, (b) means for filtering the sum dynamic equalization input signal so
5 as to generate a sum dynamic equalization output signal, and (b) means for applying the sum
6 dynamic equalization output signal to the sum signal transmission path.

1 47. An adaptive signal weighting assembly according to claim 46, further
2 characterized wherein the filtering provided by the difference dynamic equalization filter
3 means is characterized by a difference dynamic equalization phase response, the filtering
4 provided by the variable emphasis filter is characterized by a variable emphasis phase
5 response, the filtering provided by the sum dynamic equalization filter means is characterized
6 by a sum dynamic equalization phase response, the difference dynamic equalization, variable
7 emphasis, and sum dynamic equalization phase responses all varying according to a function
8 of the second control signal.

1 48. An adaptive signal weighting assembly according to claim 30, further
2 characterized further including composite modulator means for receiving the encoded





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3 difference signal and the conditioned sum signal and for generating therefrom a composite
4 modulated signal.

1 49. (New) A system according to claim 7, wherein the pilot frequency is
2 substantially 15,734 Hz.

1 50. (New) A system according to claim 7, wherein DC signal energy is absent
2 from the signal content of the information signal so as to prevent ticking.

1 51. (New) A system according to claim 8, wherein the pilot frequency is
2 substantially 15,734 Hz.

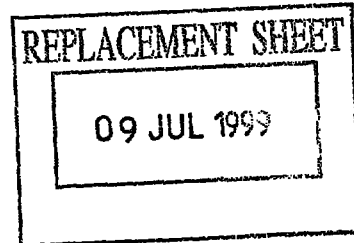
1 52. (New) A system according to claim 8, wherein DC signal energy is absent
2 from the signal content of the information signal so as to prevent ticking.

1 53. (New) A system according to claim 9, wherein the pilot frequency is
2 substantially 15,734 Hz.

1 54. (New) A system according to claim 9, wherein DC signal energy is absent
2 from the signal content of the information signal so as to prevent ticking.

1 55. (New) An adaptive digital signal weighting system for use with a digitally
2 represented difference signal representing the difference between two stereophonic audio
3 signals and a digitally represented sum signal representing the sum of the two stereophonic
4 audio signals, the system including a difference signal path for transmitting a first electrical
5 information signal containing information relating to the difference signal of a predetermined
6 bandwidth through said system, and a sum signal path for transmitting a second electrical
7 information signal containing information relating to the sum signal of a predetermined
8 bandwidth, said system further comprising:

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9 a first digital filter arrangement disposed in said difference signal path and constructed
10 and arranged so as to vary the gain impressed on the portion of said first information signal
11 within a first select spectral region within said predetermined bandwidth by a first variable
12 gain factor, said first variable gain factor varying in response to and as a function of a first
13 control signal, wherein said digital filter arrangement introduces phases errors in said first
14 information signal as a function of frequency;

15 a first control signal generator constructed and arranged so as to digitally generate said
16 first control signal only in response to and in accordance with the signal energy of said
17 digitally represented difference signal within a second select spectral region including at least
18 a part of said first select spectral region;

19 a digital gain controller, disposed in said difference signal path and coupled to said
20 digital filter arrangement, and constructed and arranged so as to vary the signal gain
21 impressed on said first information signal substantially throughout said predetermined
22 bandwidth by a second variable gain factor, said second variable gain factor varying in
23 response to and as a function of a second control signal; and

24 a second control signal generator constructed and arranged so as to digitally generate
25 said second control signal in response to and as a function of the signal energy of said
26 digitally represented difference signal substantially within a third select spectral region within
27 said predetermined bandwidth; and

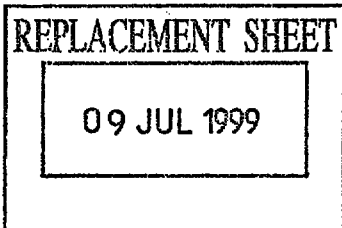
28 a signal compensation arrangement disposed in the sum signal path for compensating
29 for said phase errors introduced by said digital filter arrangement in the difference signal path;

30 wherein the digital filter arrangement, first control signal generator, digital gain
31 controller, and the second control signal generator each operate at a predetermined sample rate
32 so as to preserve the signal content of the information signal.

1 56. (New) An adaptive digital signal weighting system according to claim 55,
2 wherein the signal compensation arrangement includes a static phase equalization filter.

1 57. (New) An adaptive digital signal weighting system according to claim 55,

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wherein the signal compensation arrangement includes a dynamic phase equalization filter.

58. (New) An adaptive digital signal weighting system according to claim 55, wherein the signal compensation arrangement includes an all pass filter having a relatively flat amplitude response and a preselected phase response.

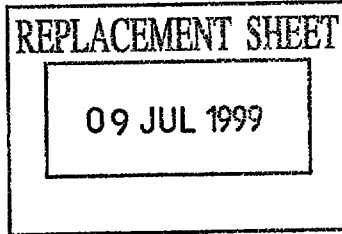
59. (New) An adaptive digital signal weighting system according to claim 55, wherein the signal compensation arrangement includes an IIR filter.

60. (New) A digital implementation of an analog adaptive digital signal weighting system for use with a digitally represented difference signal representing the difference between two stereophonic audio signals, the system including a signal path for transmitting an electrical information signal containing information relating to the difference signal of a predetermined bandwidth through said system, said analog adaptive digital weighting system being of the type including first analog control signal generator including a first analog filter component and a second analog control signal generator including a second analog filter component, said digital implementation further comprising:

a digital filter arrangement disposed in said signal path and constructed and arranged so as to vary the gain impressed on the portion of said information signal within a first select spectral region within said predetermined bandwidth by a first variable gain factor, said first variable gain factor varying in response to and as a function of a first control signal;

a first digital control signal generator constructed and arranged so as to digitally generate said first control signal only in response to and in accordance with the signal energy of said digitally represented difference signal within a second select spectral region including at least a part of said first select spectral region, said first digital control signal generator including a first digital filter component having an amplitude response such that the difference between the amplitude response of the first digital filter component and the amplitude response of the first analog filter component is minimized irrespective of their respective phase responses;

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21 a digital gain controller, disposed in said signal path and coupled to said digital filter
22 arrangement, and constructed and arranged so as to vary the signal gain impressed on said
23 information signal substantially throughout said predetermined bandwidth by a second
24 variable gain factor, said second variable gain factor varying in response to and as a function
25 of a second control signal;

26 a second digital control signal generator constructed and arranged so as to digitally
27 generate said second control signal in response to and as a function of the signal energy of
28 said digitally represented difference signal substantially within a third select spectral region
29 within said predetermined bandwidth, said second digital control signal generator including a
30 second digital filter component having an amplitude response such that the difference between
31 the amplitude response of the second digital filter component and the amplitude response of
32 the second analog filter component is minimized irrespective of their respective phase
33 responses; and

34 wherein the digital filter arrangement, first control signal generator, digital gain
35 controller, and the second control signal generator each operate at a predetermined sample rate
36 so as to preserve the signal content of the information signal.

1 61. (New) A digital implementation according to claim 60, wherein said first
2 digital filter component comprises a spectral bandpass filter.

1 62. (New) A digital implementation according to claim 60, wherein the first
2 digital filter component comprises an IIR filter.

1 63. (New) A digital implementation according to claim 60, wherein the first
2 digital filter component comprises a cascade of IIR filters.

1 64. (New) A digital implementation according to claim 60, wherein said second
2 digital filter component comprises an IIR filter.

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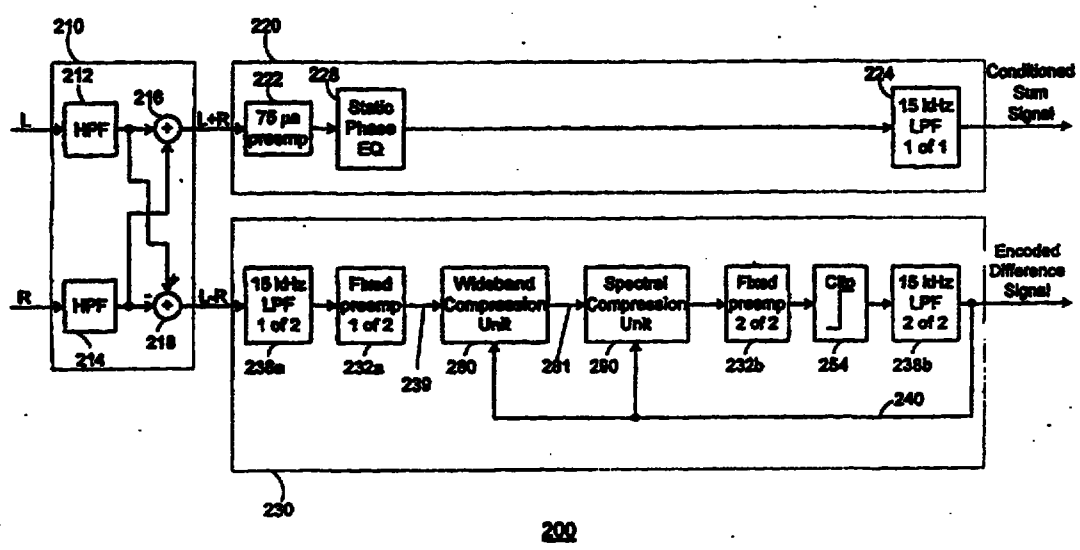
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(54) Title: **BTSC ENCODER**



(57) Abstract

A BTSC encoder (200) includes a left digital high pass filter (212) for receiving a digital left channel audio signal (L) and generating a digital left filtered signal; a right digital high pass filter (214) for receiving a digital right channel audio signal (R) and generating a digital right filtered signal; a matrix includes an adder (216) for summing the digital left and right filtered signals and generating a digital sum signal (L+R), and including a subtractor (218) for subtracting one of the digital left and right filtered signals from the other of the digital left and right filtered signals and generating a digital difference signal (L-R); a difference channel processor (230) for digitally processing the digital difference signal (L-R); and a sum channel processor (220) for digitally processing the digital sum signal (L+R).