

Description

The present invention is concerned in general with circuit arrangements which alter the dynamic range of signals, namely compressors which compress the dynamic range and expanders which expand the dynamic range. While the invention is useful for treating various types of signals, including audio signals and video (television) signals, the description of the invention is primarily in the context of the processing of audio signals. The principles of the invention may be applied to the processing of other signals by modifying the disclosed embodiments by applying known techniques. For example, compressors and expanders for video signals can act instantaneously and do not require syllabic control circuitry.

Compressors and expanders are normally used together (a compander system) to effect noise reduction; the signal is compressed before transmission or recording and expanded after reception or playback from the transmission channel. However, compressors may be used alone to reduce the dynamic range, e.g., to suit the capacity of a transmission channel, without subsequent expansion when the compressed signal is adequate for the end purpose. In addition, compressors alone are used in certain products, especially audio products which are intended only to transmit or record compressed broadcast or pre-recorded signals. Expanders alone are used in certain products, especially audio products which are intended only to receive or play back already compressed broadcast or pre-recorded signals. In certain products, a single device is often configured for switchable mode operation as a compressor to record signals and as an expander to play back compressed broadcast or pre-recorded signals.

One long sought after goal in the design of compressors, expanders and companding type noise reduction systems is a high degree of adaptiveness of the compressor and expander to applied signals. That is, the compressor, for example, ideally should provide constant gain throughout its frequency spectrum of operation except at the frequency of a dominant signal component where it would provide dynamic action according to a predetermined compression law. This goal was referred to as "conformal equalization" in GB-A-1120541. Accordingly, that document and subsequent patents derived therefrom (including US-PS 3,846,719 and US-PS 3,903,485) employed several techniques directed to achieving that goal, including, among others, techniques now commonly known as "bandsplitting" and "sliding band".

According to the bandsplitting approach, the spectrum is divided into a plurality of frequency bands, each of which is acted upon independently. In that way a dominant signal component affects dynamic action (compression or expansion) only within a portion of the overall spectrum, in contrast to a wideband approach in which dynamic action throughout the entire spectrum is affected by a dominant signal component. Thus, a bandsplitting system provides a greater degree of adaptiveness or conformance than a wideband system. In theory, a highly adaptive or conformal system could be provided by dividing the overall spectrum into a very large number of frequency bands; however, the complexity and cost of such an arrangement makes it impractical. Consequently, a design compromise is made by selecting a reasonable number of frequency bands capable of providing satisfactory performance. In one well known commercially successful bandsplitting companding type audio noise reduction system (commonly known as A-type noise reduction) four bands are employed ("An Audio Noise Reduction System", by Ray Dolby, J. Audio Eng. Soc., October 1967, Vol. 15, No. 4, pp. 383-388). However, such systems suffer from the same problems as does a wideband noise reduction system, although to a lesser degree because the band is divided up and the problems tend to be confined to the individual bands. These problems are well known in the design of noise reduction systems and include loss of noise reduction effect and the related problems of noise modulation and signal modulation at frequencies not masked by the dominant signal component when a change of gain takes place in response to a dominant signal component. Such problems are chiefly a result of a system failing to be perfectly conformant to the dominant signal. The degree to which such problems are audible also depends on how far the system departs from perfect complementarity. If, for example, the transmission channel response is irregular or unpredictable within the passband of the compressor and expander, then signal modulation effects will not be compensated in the expander.

A dominant signal component is a signal component having a substantial enough level so as to effect dynamic action within the frequency band under consideration. Under complex signal conditions there may be more than one dominant signal component or a dominant signal component and sub-dominant signal components. In a compander system which relies on complementarity of the compressor and expander, all of the signal components must be compressed and expanded in accordance with a defined compression/expansion law in order that the signal spectrum including the dominant signal component (and other signals affected by dynamic action) can be restored to their correct levels in the expander. This requirement excludes the usefulness in compander systems of various known adaptive and tracking filter techniques and so-called "single ended" noise reduction systems (which operate only on a reproduced signal) in which the filter action is not subject to predetermined compression/expansion laws and whose action may be

unpredictable in the presence of multiple signals.

Another approach useful in working toward the goal of increased adaptiveness or conformance is the sliding band technique, which employs signal dependent variable filtering to achieve limiting. Generally, a dominant signal component causes the cutoff or turnover frequency (or frequencies) of one or more variable filters (e.g., high pass, low pass, shelf, notch, etc.) to shift so as to compress or expand the dominant signal component.

A sliding band system operating only in a single high frequency band is described in US-PS Re 28,426 and US-PS 4,490,691. This system, which forms the basis for the well known consumer companding type audio noise reduction system known as B-type noise reduction, includes, in a dual path arrangement, a side path having a fixed high pass filter in series with a variable filter.

A "dual path" arrangement is one in which a compression or expansion characteristic is achieved through the use of a main path which is essentially free of dynamic action and one or more secondary or side paths having dynamic action. The side path or paths take their input from the input or output of the main path and their output or outputs are additively or subtractively combined with the main path in order to provide compression or expansion. Generally, a side path provides a type of limiting or variable attenuation and the manner in which it is connected to the main path determines if it boosts (to provide compression) or bucks (to provide expansion) the main path signal components. Such dual path arrangements are described in detail in US-PS 3,846,719; US-PS 3,903,485; US-PS 4,490,691 and US-PS Re 28,426.

A high frequency variable shelving filter in a single path arrangement (e.g., the dynamic action is accomplished in a single signal path) for a companding audio noise reduction system is set forth in US-PS 3,911,371. In the embodiments of Figures 1 and 2 of US-PS 3,665,345 a dual path arrangement is set forth in which the side path comprises a variable shelving filter having an all-pass characteristic in its quiescent condition. Another approach for providing a variable shelving response for compander systems is set forth in US-PS 3,934,190.

One drawback of these sliding band arrangements is that in the presence of a dominant high frequency signal component the variable filter turnover frequency shifts to a frequency above that signal component thereby restricting the frequency area at lower frequencies in which noise reduction is provided. The loss of noise reduction may be more noticeable audibly than in bandsplitting systems and the related side effects (noise modulation and signal modulation) may be more severe than in fixed band arrangements because of a multiplication effect that is inherent in sliding band systems. This effect results from the way in which sliding band systems provide compression. If, for example, there is a dominant high frequency signal and 2 dB of gain reduction is required at that frequency, the variable filter cutoff frequency should shift to the extent necessary to provide that amount of attenuation along the filter slope. However, for lower frequencies, further removed from the new filter cutoff frequency, the effect may be 5 or 10 dB of dynamic action, for example, with a consequent loss of all or most of the noise reduction-effect along with possible audible signal or noise modulation. In other words, in this example, a 2 dB change in a dominant signal can cause a 5 or 10 dB change in gain at frequencies removed from the dominant signal. Figure 1 is an idealized compressor characteristic response curve illustrating this effect. (Throughout this document the characteristic response curves illustrated in the various Figures are those of compressors, it being understood that the respective expander characteristic is the complement of the compressor characteristic.) Under relatively rare conditions, when very high frequency dominant signal components (cymbals, for example) control the sliding band filter, there may be audible modulation of non-dominant mid-band signal components that are also present if the expander does not properly track the compressor. This problem is called the "mid-band modulation effect". One approach in solving the problem is set forth in said US-PS 4,490,691.

In a fixed band arrangement the same amount of gain reduction would occur throughout the frequency band (whether wide band or one frequency band of a bandsplitting system) in response to a dominant signal component. Thus, while signal or noise modulation may occur, there is no multiplication of the effect: a 2 dB change in the level of a dominant signal component would cause no more than a 2 dB change in gain at frequencies removed from the dominant signal component. However, viewed from the standpoint of noise reduction effect this is a disadvantage of a fixed band arrangement--the full noise reduction effect is not obtained anywhere within the frequency band of operation when limiting occurs in response to a dominant signal component. Figure 2 is an idealized compressor characteristic response curve illustrating this effect. Although it is not multiplied, there is also the potential for noise and signal modulation throughout the entire frequency band in which the fixed band action occurs.

Despite the disadvantages mentioned, an advantage of a sliding band arrangement is that the full noise reduction effect is obtained at frequencies above the dominant signal component (or below the dominant signal component in the case of a sliding band system acting downward in frequency). Thus an

arrangement that achieves the advantages of fixed band and sliding band systems (e.g., the advantage of fixed band is that there is no multiplication of modulation effects and the advantage of sliding band is that there is minimum signal or noise modulation above the dominant signal frequency) without the disadvantages of each (e.g., the disadvantage of fixed band is noise and signal modulation throughout its operating range—although not multiplied and the disadvantage of sliding band is the mid-band modulation effect) would be desirable. The invention provides this combination.

In a broader sense the invention as claimed is intended to remedy the above mentioned drawbacks of prior art dynamic range modifying circuits. It solves the problem of how to design such circuit that the advantages of various dynamic and passive characteristics are obtained in order to achieve compressor, expander and compander systems having a high degree of conformance or adaptation to dominant signals without undue circuit complexity.

Summary of the Invention

The present invention is based on the recognition that the ideal of conformal equalization can be more closely approached by compressor, expander and compander type noise reduction arrangements in which a plurality of compression/expansion/equalization characteristics are superposed or overlaid upon one another in such a way that one or more of the characteristics is hidden or concealed until, as dominant signal components appear, the hidden characteristics are revealed and become active. Thus, according to the invention, the quiescent characteristic, which provides a defining umbrella or envelope that conceals one or more latent characteristics, is modified so that the latent characteristic or characteristics emerge in response to dominant signal components in order to provide a more effective adaptive equalization than provided by prior art circuit arrangements.

This unveiling of characteristics can be described as "action substitution" in the sense that the action resulting from one (or perhaps more than one) characteristic is substituted for one or more other characteristic actions that have the potential to operate in the same frequency and level regions when the level and spectral content of the input signal components change. Preferably, the substitution is such that, with respect to any non-dominant signal components, the transmission is maximized in the compressor and minimized in the expander. The invention allows the designer of compressors, expanders and companding noise reduction systems greater flexibility in building up response characteristics without resorting to impractical circuit complexity. The invention provides compressors, expanders and noise reduction compander systems having improved abilities to discriminate among dominant and non-dominant signal components and to confine dynamic action to dominant signal components only. By providing a noise reduction encoder (compressor) which essentially maintains a constant boost except where there is a dominant signal component, the noise reduction decoder (expander) has a very stable noise floor, which is essential to a high quality noise reduction system.

In order to best implement superposition according to this invention an arrangement is required such that substantially full "action substitution" is obtained. In an action substitution arrangement, characteristics dynamically emerge and appear throughout all or part of the spectrum depending on the level and spectral content of dominant signal components. Action substitution provides for the revealing of latent characteristics. Ideally, in a superposition arrangement, a characteristic appears only if it can provide an optimal effect within all or part of the spectrum; that characteristic is then fully effective within that region. For example, in the case of a compressor, for non-dominant signal components, the ideal characteristic is that which provides the highest signal output; or, in the case of an expander, the lowest signal output. Each characteristic is substantially fully effective for non-dominant signal components except in the crossover region of effectiveness of the characteristics; in the crossover region, the overall combined characteristic does not exceed the maximum action of any of the characteristics acting alone. The maximum overall output is not greater than the maximum output of any single characteristic. The characteristics are hidden by one another, the one or ones exposed depending on which has the greatest effect at any particular frequency under the signal conditions at that instant.

For discussion purposes let the gains in a compressor system be arranged such that extremely low level signals pass through the apparatus without attenuation. That is, the maximum possible action is that of providing a gain of unity; somehow to achieve this gain over as broad a range of frequencies as possible, in the presence of higher level (dominant) signals, is the task of the system.

Thus, in a superposed action compressor, the output of the first filter and compressor element is noted. The output represents the completed part of the total potential action. The uncompleted part of the action is the input minus the completed part. The uncompleted part is therefore derived and fed into the next element to see what it can contribute (it would be wrong to feed the whole input signal into the second

element, because this might result in double action). The output of the second element is the contribution of this particular element to the total action; its action (hopefully, more successful) augments the partially inadequate action of the first element. In an extreme case, in which the output of the first element may be negligible at a particular frequency, the action of the second element is effectively substituted for that of the first.

Any remaining potential action is then the input signal minus the outputs of the first two elements. Thus this difference signal is fed to the third element, etc. If there are enough elements, at least some of which have sufficient discrimination against the dominant signal(s) present, then at non-dominant frequencies the total of all the element gains will approach the desired sum of unity.

At the dominant frequency the gain of the overall compressor cannot be allowed to be unity but must obey a specified compression law in order to avoid channel overload and to permit restoration of the signal by the expander.

Superposition effects in accordance with the invention are achieved by various circuit arrangements, the most basic being those which follow directly from the above presentation—that is, particularly those in which the effective result is to feed the inputs to a plurality of elements differentially and to sum the element outputs. For example, if each element has an input terminal, a reference potential (e.g., ground) terminal, and an output terminal, then the input to the first element is the input signal with respect to the reference potential, the input to the second element is the input signal with respect to the output of the first element (e.g., the difference between the input signal and the output of the first element), the input to the third element (if there is a third element) is the input signal with respect to the sum of the outputs of the first and second elements (e.g., the difference between the input signal and the sum of the outputs of the first and second elements), and so on if there are more elements. The outputs of the elements are summed to provide the overall output. This can be implemented in a number of substantially equivalent ways including a circuit topology such as in Figure 3 in which the inputs and outputs are literally taken as just described.

However, there is a simpler way, shown in Figure 4, in which the three-terminal network elements are linked together with less complexity and yet achieve the same results as the arrangement of Figure 3. The input signal is applied to the input terminal of each element. The first element input is with respect to a reference potential (e.g., its second terminal is connected to a reference potential); the second element input is with respect to the output of the first element (e.g., its second terminal is connected to the output terminal of the first element rather than to the reference potential); the third element input (if there is a third element) is with respect to the output of the second element (e.g., its second terminal is connected to the output element of the second element rather than to the reference potential); and so on if there are more elements. The overall output is taken at the output terminal of the last element. This latter arrangement, which may be referred to as a "stack" of elements, is the preferred way in which to practice the invention because of its simplicity. However, other circuit topologies can provide the same or similar results, although with greater complexity and cost of implementation. For example, Figure 5 shows another arrangement which is useful in demonstrating the transition from the Figure 3 topology to the Figure 4 topology. Figures 3, 4 and 5 are equivalent to each other: Figure 5 is a transition between Figures 3 and 4, retaining the summed outputs while removing the second terminals of all but one element from the reference potential.

Although the invention is applicable generally to combinations of elements having various dynamic and passive characteristics, a very useful combination of characteristics in practice and for the purposes of explanation is the superposition of a fixed band dynamic characteristic and a sliding band dynamic characteristic. By applying the teachings of the present invention the advantages of both characteristics can be obtained while avoiding their disadvantages. Thus, if a sliding band characteristic and a fixed band characteristic are superposed in substantially the same frequency range (wide band or a defined band) and level range, the quiescent characteristic of the superposed combination appears the same as the quiescent characteristic of either one taken alone because the two quiescent characteristics are the same. When a dominant signal component appears within their frequency range each characteristic reacts—the fixed band characteristic drops uniformly in level across the frequency range similar to the way it would if acting by itself and the sliding band characteristic slides similarly to the way it would if acting by itself.

However, the two actions are no longer independent: to some extent each acts with reference to the other. When these changes occur the two characteristics, which appeared as one characteristic in the quiescent condition (Figure 6A), are now revealed: the combined characteristic appears as that of a sliding band characteristic above (or below, depending on whether the sliding band acts upwardly or downwardly in frequency) the frequency of the dominant signal and it appears as a fixed band characteristic below (or above) the frequency of the dominant signal. Figure 6B shows an example in which the sliding band is above the dominant signal and Figure 6C shows an example in which the sliding band is below the dominant signal. Two regimes of operation are revealed, divided at the frequency of the dominant signal.

Thus, the region which the sliding band characteristic would have left "uncovered" is supplemented by the fixed band characteristic which, in effect, provides a floor or foundation level. In other words, there is a substitution of action in response to the dominant signal component. The result is to obtain the advantages of both fixed band and sliding band arrangements while avoiding their disadvantages. Maximum noise reduction effect and minimum modulation effects are obtained above (or below) the dominant signal where the sliding band characteristic operates while avoiding the loss of noise reduction and the mid-band modulation effect below (or above) the dominant signal by the presence of the fixed band characteristic. Thus, there is no multiplication effect below (or above) the dominant frequency as would occur if the sliding band characteristic were operating alone, while obtaining the advantages of the sliding band characteristic above (or below) the dominant frequency.

An even more adaptive arrangement can be achieved by providing a bandsplitting arrangement in which the high frequency band and low frequency band each comprise superposed fixed band/sliding band characteristics. In the high frequency band the sliding band acts upwardly in frequency while in the low frequency band the sliding band acts downwardly in frequency. In the quiescent condition the characteristics overlap so as to provide a flat overall characteristic. By choosing gentle filter slopes (say, 6 dB/octave) and a common quiescent corner frequency in the middle of the frequency band (say, 800 Hz for an audio system), excellent tracking of a dominant signal by both the high and low frequency bands throughout a substantial portion of the band under processing is possible. The quiescent response of such an arrangement, in which the high frequency and low frequency bands have the same low level gain, is flat as shown in Figure 7A.

Figure 7B shows the tracking of a single high-level dominant signal at various frequencies for a multi-stage compressor arrangement such as described in greater detail below in connection with the description of Figures 22, 23, and 24 (in each case, without the respective spectral skewing and antisaturation networks). Because the compressor arrangements of Figures 22, 23, and 24 employ two low-frequency stages having a low level gain of 16 dB and three high-frequency stages having a low level gain of 24 dB, the quiescent response, as shown in Figure 7B, is 16 dB at low frequencies, rising to 24 dB at high frequencies. In the example of Figure 7B, for each high-level (0 dB relative to a reference level which is taken to be about 20 dB below the maximum level in the system) tone (a dominant signal) at various frequencies (100 Hz, 200 Hz, 400 Hz, 800 Hz, 1.6 kHz, 3 kHz, and 6 kHz), a low-level tone (-60 dB to -70 dB) is swept through the spectrum to show the overall response in the presence of the dominant signal. A sliding band response appears above and below the dominant signal. The presence of two dominant signals results in a fixed band response between the dominant signals and sliding band response at frequencies above and below their frequencies where maximum noise reduction is most critical (Figure 7C; in Figure 7C as in Figure 7A, the high frequency and low frequency bands have the same low level gain).

The good tracking behavior exhibited in Figures 7B and 7C is made possible by the use of both fixed and sliding band characteristics in each of the two bands of the bandsplitting arrangement. For example, in Figure 7C there would clearly be a serious deficiency of non-dominant signals (poor noise reduction effect) in the region between the two major dominant signals if the two fixed bands were not used. If desired, a mid-band (a pass band, 400 Hz to 1.6 kHz, for example), overlapping in part the high and low frequency bands, could also be incorporated to improve the noise reduction in the mid-frequency range when both extremely low frequency and extremely high frequency dominant signals are simultaneously present. The output of the mid-band element could be fed to the bottom of both high and low frequency stacks. This latter refinement does not appear to be necessary in practical audio noise reduction circuits.

Note that action substitution makes good signal tracking possible. Consider the use of conventional high slope (12 dB/octave or more) high and low pass variable filters. Firstly, the filter outputs would not combine to a flat characteristic in the quiescent condition; using a similar attenuation shape for both the low and high frequencies, this is possible only with single pole filters (Figure 7A). Secondly, as soon as one or both filters slide, a non-dominant signal deficiency would appear (poor noise reduction).

In the present invention the filters defining the stage boundaries are single pole filters for optimal low-level response. In the presence of signals, additional single pole sliding band shelving filters are used, whereby a quasi-two-pole result is achieved without the excessive phase shift (phase reversal) inherent in two-pole filters. The intervention of the fixed band, via action substitution, further tailors the response. The use of several staggered-level stages then provides a further quasi-multi-pole effect to compound the resultant responses--i.e., the effective network steepness or ability of the system to discriminate between dominant and non-dominant signal components.

Regarding dynamic recovery characteristics of the overall compressor or expander, action substitution can further contribute to improved performance. A fixed band will have a recovery time that is essentially independent of frequency, at least in the pass band. A sliding band circuit has a fast recovery time for non-

dominant signals at the pass band end of the spectrum, and a slow recovery time for non-dominant signals at the stop band end of the spectrum. The choice of control circuit recovery times is therefore a matter of compromise between this recovery time situation and the amount of steady state and modulation distortion obtained. The compromise is much easier, however, with the use of the action substitution technique. In particular, the fixed band provides a definite and rapid recovery time for the overall system, so that the sliding band can employ longer time constants than would otherwise be desirable. This results both in low modulation distortion and a fast recovery time.

Brief Description of the Drawings

Figure 1 is an idealized compressor characteristic response curve illustrating the prior art sliding band multiplication effect.

Figure 2 is an idealized compressor characteristic response curve illustrating the prior art fixed band limiting effect.

Figure 3 is a block diagram illustrating generally a circuit topology useful in practising the invention.

Figure 4 is a block diagram illustrating generally a circuit topology equivalent to the topology of Figure 3 and preferred for practising the invention.

Figure 5 is a block diagram of a further circuit topology and one which is useful in demonstrating the transition from the topology of Figure 3 to that of Figure 4.

Figure 6A is an idealized compressor characteristic response curve showing the quiescent response of fixed band and sliding band elements superposed in accordance with the invention.

Figure 6B is an idealized compressor characteristic response curve showing the response slightly above their thresholds of fixed band and sliding band elements superposed in accordance with the invention, the sliding band acting upward in frequency.

Figure 6C is an idealized compressor characteristic response curve showing the response slightly above their thresholds of fixed band and sliding band elements superposed in accordance with the invention, the sliding band acting downward in frequency.

Figure 7A is an idealized compressor characteristic response curve showing the quiescent response of high frequency and low frequency circuits, each having fixed band and sliding band elements superposed in accordance with the invention, the circuits having a common corner frequency of 800 Hz.

Figure 7B is an idealized compressor characteristic response curve showing the tracking response of the same high frequency and low frequency circuits referred to in connection with Figure 7A in the presence of a dominant signal at various frequencies.

Figure 7C is an idealized compressor characteristic response curve showing the response of the same high frequency and low frequency circuits referred to in connection with Figure 7A in the presence of two dominant signals.

Figure 8 is a schematic circuit diagram useful in explaining the basic operation of the preferred embodiment of the invention.

Figure 9 is a block diagram showing an equivalent arrangement for practicing the invention.

Figure 10 is a block diagram illustrating a modified arrangement for practicing the invention.

Figure 11 is a partially schematic block diagram demonstrating the way in which multiple characteristic actions may be linked together in accordance with the invention.

Figure 12A is an idealized compressor characteristic response curve useful in understanding the operation of the arrangement of Figure 11.

Figure 12B is a further idealized compressor characteristic response curve useful in understanding the operation of the arrangement of Figure 11.

Figure 13 is a partially schematic block diagram showing an embodiment of the invention in which a fixed band element and a sliding band element are stacked together, each having separate control circuits, with an optional cross-coupling from the control circuit of one element to another element.

Figure 14 is a modification of the embodiment of Figure 13 in which a filter is included in the coupling provided from the control circuit of one element to another element.

Figure 15 is a modification of the embodiment of Figure 13 in which a common control circuit is provided for the two elements.

Figure 16A is modification of the embodiment of Figure 14 in which the fixed band element is parked at a level other than its normal quiescent level.

Figure 16B is further modification of the embodiment of Figure 14 in which the fixed band element is parked at a level other than its normal quiescent level and in which a type of cross-coupling is provided such that the corner frequency of the sliding band element is parked at a frequency other than its normal

quiescent frequency. The figure also shows another aspect of the invention, a shunt for the gyrator inductor element.

Figure 17A is an idealized compressor characteristic response curve showing the quiescent response of a circuit of the type shown in the arrangements of Figures 13 through 16.

5 Figure 17B is an idealized compressor characteristic response curve showing the response slightly above the thresholds of the elements of a circuit of the type shown in the arrangements of Figures 13 through 16.

Figure 17C is an idealized compressor characteristic response curve showing the response at an even greater level above the thresholds of the elements of a circuit of the type shown in the arrangements of 10 Figures 13 through 16.

Figure 18 is a partially schematic block diagram of an embodiment of a high frequency fixed band/sliding band stage according to the invention.

Figure 19 is a partially schematic block diagram of an embodiment of a low frequency fixed band/sliding band stage according to the invention.

15 Figure 20 is a block diagram showing a Type I dual path arrangement employing high frequency and low frequency stages of the type described in connection with Figures 18 and 19.

Figure 21 is a block diagram showing a Type II dual path arrangement employing high frequency and low frequency stages of the type described in connection with Figures 18 and 19.

Figure 22 is a block diagram of a compander system having series staggered stages employing high 20 frequency and low frequency circuits of the type described in connection with Figures 18 and 19.

Figure 23 is a partially schematic block diagram of a switchable compressor/expander employing high frequency and low frequency circuits of the type described in connection with Figures 18 and 19 and illustrating the use of a secondary main path in order to simplify switching.

Figure 24 is a partially schematic block diagram of a switchable compressor/expander employing high 25 frequency and low frequency circuits of the type described in connection with Figures 18 and 19 and illustrating the use of an inverting amplifier in parallel with the compressor portion of the arrangement of Figure 22 in order to simplify switching.

Figure 25 is a block diagram of a prior art switching arrangement useful in explaining the operation of Figure 24.

30 Figure 26 is a characteristic compression response curve relating to the arrangements of Figures 22, 23, and 24.

Detailed Description of the Invention

35 In order to better understand the operation of the basic topology of the preferred embodiment of the present invention it is useful first to simplify the arrangement so that there are only two elements, each of which is a wideband (frequency independent) resistive attenuator as in Figure 8. The maximum possible action of each element is to provide a gain of unity. R_1 and R_2 comprise the first attenuator (R_2 being variable), a three-terminal network, receiving V_{in} at its input, having its second terminal connected to ground 40 and its output applied to the second terminal of the other attenuator via a unity-gain buffer B. The second attenuator, another three-terminal network, receiving V_{in} at its input and providing the overall combination output at its output, is comprised by R_3 and R_4 , the latter being variable.

First, assume that R_2 and R_4 are very large; then neither network provides any attenuation (viewed from their common inputs to R_1 and R_3) and V_{in} is present at nodes n_1 and n_2 (the junctions of R_1/R_2 and R_3/R_4 , 45 respectively). It follows that V_{out} must equal V_{in} because the potential across R_4 is zero and no current flows in R_4 . It is known that in a passive three terminal network, if there is no internal connection to ground, the voltage transfer function of one input to the output is the complement of the other input to the output (e.g., if the transfer function with respect to one input is "t", the transfer function with respect to the other output is "1-t", its complement). Thus, the transfer function of the top network is 1 with respect to the signal applied 50 to R_3 and its complementary transfer function with respect to the signal applied to R_4 is zero. The sum of the voltages resulting from the two transfer functions is therefore V_{in} .

Next, assume that R_2 is very large but that R_4 is very small. Thus the lower element has no attenuation and the upper element has maximum attenuation (viewed from their common inputs to R_1 and R_3). Thus, V_{in} is present at node n_1 and must also be present at node n_2 because R_4 is very small. There is no 55 contribution from the input via R_3 because the attenuation of the upper network is maximum. Thus the output is V_{in} . Or, viewed from the standpoint of transfer functions, the transfer function of the top network is zero with respect to the signal applied to R_3 and its complementary transfer function with respect to the signal applied to R_4 is one. The sum of the voltages resulting from the two transfer function is therefore V_{in} .

It is useful to demonstrate that for this general case that the reverse positioning of the elements is immaterial to the result. Thus assume that R_4 is very large but that R_2 is very small. Thus the upper element has no attenuation and the lower element has maximum attenuation (viewed from their common inputs to R_1 and R_3). Therefore, the voltage at node n_2 is V_{in} and must also be the output voltage V_{out} because R_4 is large and no current flows through it. Considering transfer functions, the transfer function of the top network is one with respect to the signal applied to R_3 and its complementary transfer function with respect to the signal applied to R_4 is zero. The sum of the voltages resulting from the two transfer function is therefore V_{in} .

Thus, if both of the elements have no attenuation (as viewed from the inputs to R_1 and R_3), then the output is V_{in} . It can be observed from the above that the output is also V_{in} if either of the elements has no attenuation (as viewed from the inputs to R_1 and R_3) regardless of the attenuation, if any, of the other element.

This analysis extends on a steady-state basis to the more complex cases as described hereinafter where attenuation is provided using field effect transistors (FETs) as the variable resistance elements and the attenuation is effective only in a portion of the frequency band of the input signal.

Several general observations can be drawn regarding the operation of the present invention. These observations are believed to be valid not only for the simple circuit arrangements discussed above, but also under most signal conditions in more complex circuit arrangements contemplated by the invention. First, the highest transmission of any element is the output. Second, the characteristics of the elements do not produce an output characteristic any greater in magnitude than the greatest magnitude of either element (e.g., the characteristics cannot provide too much effect) and third, the characteristic of one element is hidden by that of another, depending on which is greater in magnitude. While it is possible to achieve such results using other circuit arrangements, the circuit arrangements according to the present invention are preferred because they are simple to implement and they provide the desired results.

In practice, the first observation is particularly useful in the case where all but one element is knocked out in response to signal conditions, leaving only one element to provide compression or expansion. In that situation, the remaining element provides its full effect (which may be less than its maximum possible effect, depending on how it is affected by the dominant signal). An example of this would be an arrangement of a sliding band element and a fixed band element operating in the same frequency range (having responses as, for example, in Figures 6B or 6C). A dominant signal could cause the sliding band element to slide upward (or downward) so as to totally knock out its effect at or near the frequency of the dominant signal. However, the fixed band element would remain fully effective (although perhaps less than its maximum effect as a result of the presence of the dominant signal). This is the case at frequencies well below the dominant signal in Figure 6B and well above the dominant signal in Figure 6C. However, it will be noted that in the crossover region (at the frequency of the dominant signal in Figures 6B and 6C), where the overall response changes from the fixed band to the sliding band response, there is a slight augmentation in the response such that the overall response is somewhat greater than the response of either element. This augmentation effect occurs in general where the operating regimes of characteristics operating at less than their maximum effect crossover. The effect is reflected in the presence of the product term of Equation 1.

The second observation is fundamental to the "defining umbrella" effect of the invention. That is, the maximum possible effect of the combination of characteristics is no greater than the greatest effect of each characteristic. However, when characteristics operate at less than their maximum effect, the above noted augmentation of responses occurs in crossover regions.

The third observation is also basic to the invention, permitting the substitution of characteristic actions in response to changes in signal conditions.

The invention thus provides a circuit for modifying the dynamic range of input signal components within a frequency band, in which a plurality of circuit elements are interconnected, each having its own passive or dynamic characteristic action, at least one having a dynamic characteristic action, the characteristic actions operating at least partially within substantially the same frequency and level region, the circuit elements being interconnected such that for some level and spectral content of the input signal components the characteristic action of one circuit element substitutes for the characteristic action of another circuit element in at least a portion of the frequency band. The interconnection of the circuit elements provides for characteristic action substitution by superposition of characteristics wherein an overall circuit characteristic action is derived from the individual characteristic actions of the plurality of circuit elements such that for any combination of level and spectral content of the input signal components an individual characteristic action defines the overall frequency band or individual characteristic actions each define portions of the overall frequency band, the individual characteristic action or characteristic actions which define all or a portion of the overall frequency band suppressing the effect of any other individual characteristic action or

characteristic actions having less effect within the same frequency range. The overall characteristic action for any combination of level and spectral content of input signal components has substantially no greater an effect at any frequency than any individual characteristic operable at that frequency.

It can be shown that the overall voltage transfer function of the combination of two elements interconnected as described is of the form

$$V_{out} = V_{in}[t_1(s) + t_2(s) - t_1(s)t_2(s)] \quad (\text{Eqn. 1}),$$

where V_{in} is the applied voltage, V_{out} is the output voltage, $t_1(s)$ and $t_2(s)$ are the transfer functions of each element.

This equation confirms the operation as described above. Namely, the overall output is the sum of the transfer functions but subtracted therefrom is their product. In other words, to the extent that the transfer functions may overlap, a factor is subtracted from the sum of the transfer functions. The presence of this third term is essential to the action substitution and superposition effects of the invention.

In terms of the fixed band/sliding band example, the fixed band and sliding band responses maintain their respective characteristics where they do not overlap; within the area of overlapping (e.g., their crossover region of effectiveness) they affect each other so that the overall result in the area of overlap is substantially no greater than either of the elements acting alone.

The same basic principles apply to larger numbers of elements connected generally in the same manner as described, as set forth in connection with Figures 3 through 5 and 8 through 11, although the transfer function equations become more complex. Larger combinations of elements can be analyzed by an iterative process, extrapolating from the above two element analysis: for example, for a combination of three elements first consider the first two elements, applying their output as though it were the output of a single element in a two element combination. It can be shown that the overall transfer function of the combination of three elements interconnected in the manner described is of the form

$$V_{out} = V_{in}[t_1(s) + t_2(s) + t_3(s) - t_1(s)t_2(s) - t_1(s)t_3(s) - t_2(s)t_3(s) + t_1(s)t_2(s)t_3(s)] \quad (\text{Eqn. 2}),$$

where V_{in} is the applied voltage, V_{out} is the output voltage, $t_1(s)$, $t_2(s)$ and $t_3(s)$ are the transfer functions of each element. The same general pattern is present as with two elements with the exception of the presence of a further term, which is a product of the three transfer functions.

As a practical matter, the circuits may include both gains and attenuations. The unity gain buffer between elements is symbolic and can be omitted if the impedances are suitable, provided that loading effects are avoided. In the simplest case of the Figure 8 arrangement the buffer B might be omitted and a direct connection made between node n_1 and variable resistor R_4 if the impedances are appropriate. While the resulting circuit may have interactions between the elements that may not occur when a buffer is used, the circuit performance may be acceptable for certain applications.

The above equation number 1 describing the circuit of Figure 8 suggests an equivalent arrangement shown in block diagram form in Figure 9. The input signal is applied to three paths whose outputs are summed. The first path contains the transfer function $t_1(s)$ in block 2, the second path the transfer function $t_2(s)$ in block 4, and the third path the series combination of transfer functions $t_1(s)$ and $t_2(s)$, shown in blocks 6 and 8, respectively. The product of the transfer functions in the third path is inverted by inverter 10 so as to subtract when summed in block 12. While such a configuration would be impractical because of its additional complexity, it nevertheless illustrates that the same results can be achieved using alternative circuit arrangements. Equation 2 suggests a similar modified configuration for three elements. The teachings of the present invention could also be practiced by a special purpose or general purpose digital computer controlled by custom software programming which implements an algorithm practicing the action substitution and superposition principles of the invention.

Figure 3 illustrates the equivalent arrangement described in the summary of the invention. Four three-terminal elements 14, 16, 18 and 20 are shown having transfer functions $t_1(s)$, $t_2(s)$, $t_3(s)$, and $t_4(s)$, respectively. Each element has its second terminal connected to ground and has its third terminal output connected to a summing means 22. The input signal V_{in} is applied to the first terminal of element 14 with respect to ground and to a first summing means 24. The input to element 16 is the output of element 14 subtracted by summing means 24 from the input V_{in} . The input to element 18 is the output of element 16 subtracted by summing means 26 from the input to element 16. The input to element 20 is the output of element 18 subtracted by summing means 28 from the input to element 18.

Figure 4, also referred to in the summary of the invention, illustrates more generally the preferred embodiment discussed in connection with Figure 8. In the arrangement of Figure 4 four three-terminal

elements 30, 32, 34, and 36 are shown having transfer functions $t_1(s)$, $t_2(s)$, $t_3(s)$, and $t_4(s)$, respectively. The input signal V_{in} is applied to the first terminal of elements 30, 32, 34, and 36 with respect to the reference potential (shown as ground). Only the second terminal of element 30 is connected to a reference potential. The input to the second terminal of element 32 is the output of element 30 (from its third terminal). The input to the second terminal of element 34 is the output of element 32 (from its third terminal). The input to the second terminal of element 36 is the output of element 34 (from its third terminal). The overall output is taken between the third terminal of element 36 and the reference potential. This general arrangement can be described as "stacking" of elements and is believed to be the simplest and most effective way in which to implement the invention.

Figure 5, the transition between the arrangements of Figures 3 and 4 referred to in the summary of the invention, illustrates another possible circuit topology that is equivalent. Four elements are used solely for the purposes of illustration in Figures 3, 4, and 5. Superposition effects in accordance with the teachings of this invention can be achieved using two or more elements, at least one being dynamic, as is explained further below. In the arrangement of Figure 5 four three-terminal elements 38, 40, 42, and 44 are shown having transfer functions $t_1(s)$, $t_2(s)$, $t_3(s)$, and $t_4(s)$, respectively. The input signal V_{in} is applied to the first terminal of elements 38, 40, 42, and 44 with respect to the reference potential (shown as ground). Only the second terminal of element 38 is connected to the reference potential. The input to the second terminal of element 40 is the output of element 38 (from its third terminal). The input to the second terminal of element 42 is the summation (from summing means 46) of the output of element 38 and the output of element 38 subtracted from the output of element 40 (derived in summing means 48). The input to the second terminal of element 44 is the summation (from summing means 50) of the output of summing means 46 and the output of summing means 48 subtracted from the output of element 42 (derived in summing means 52). The overall output is taken from summing means 54 which combines the output from element 38, summing means 48, summing means 52, and summing means 56, which subtracts the output of summing means 50 from the output of element 44.

The stacking arrangement of Figure 4 can be further modified by introducing branching such as shown in Figure 10. Such arrangements may be useful in creating more complex superposition effects than are possible with a direct or straight linkage such as in Figure 4. In the Figure 10 modification the second "level" in the stack has two elements rather than one. Thus, the input V_{in} is applied to the first terminal of blocks 58, 60, 62 and 66 having respective transfer functions $t_1(s)$, $t_2(s)$, $t_3(s)$ and $t_4(s)$. The second terminal of the three-terminal network of block 58 is connected to a reference potential (e.g., ground) and its third terminal feeds the second terminals of blocks 60 and 62. The outputs from the third terminals of blocks 60 and 62 are added in summing means 64 and applied to the second terminal of block 66. The overall output is taken from the third terminal of block 66. Other branching arrangements are possible: the arrangement of Figure 10 is only one example of the manner in which elements may be configured in accordance with the invention.

Although not shown in Figures 3, 4, 5, and 10, the linkages between elements may be buffered or unbuffered, but preferably are buffered, as discussed above. In addition, while the general arrangements thus far shown employ three-terminal networks, equivalent circuit arrangements are possible based on these teachings in which some or all of the elements are three or four terminal networks each of which "floats" or is connected to a reference potential. For example, transformers and various isolation techniques could be employed to achieve the basic "differential feed, summed output" arrangement inherent in Figures 3 through 5 and 8 through 10.

With respect to any combination of elements in accordance with the teachings of the invention, in order to provide an overall compressor or expander action, at least one element must have a transfer function that is a compressor or expander function. All of the elements which are active (e.g., have characteristics that vary dynamically in response to signal conditions) generally must act in the same sense (e.g., provide compressor or expander action). One or more elements may have a passive characteristic (e.g., the characteristic does not vary dynamically in response to signal conditions). It may be useful to have more than one passive element provided that the active element has a dynamic characteristic such that as it varies different aspects of the passive characteristics are revealed.

Although arrangements of elements in accordance with the invention can function as stand alone compressors and expanders, it is preferred to employ them in one or more side paths of dual path compressor and expander arrangements of the type generally described in US-PS 3,846,719, US-PS 3,903,485, US-PS 4,490,691; and US-PS Re 28,426.

Active elements each preferably have their variable conditions controlled by their own control circuits, although for certain purposes it may be acceptable to control more than one active element with a single control circuit. Where individual control circuits are employed, although it is possible for these control

circuits to operate independently only in response to the signals in each element, in practice the overall circuit performance may be improved by applying additional signals to the control circuits, as described further below. It will be apparent that the simple analysis based on steady-state attenuator elements may not fully apply in practice under dynamic conditions where the elements have control circuits and where those control circuits can have interconnections to other elements and portions of the overall system environment.

It will also be appreciated that some active elements may have thresholds at which dynamic action will begin in response to a dominant signal. In designing a stack of elements, the selection of different threshold levels among the active elements may be advantageous so as to influence the sequence in which the characteristics of the various elements emerge in response to various dominant signal conditions. In general, however, in order to achieve superposition effects, the active or dynamic elements must be active in generally the same level range. That is, the range of input signal levels that cause dynamic action should be substantially the same, subject to variations of a few decibels.

The arrangement of Figure 11 is useful in illustrating the way in which multiple elements with various characteristics may be usefully combined and in describing the setting of relative thresholds among elements in a stack. For simplicity and ease of understanding only the principal components necessary to illustrate the concept are shown in this Figure (and the following Figures 13 through 16). The arrangement may form the noise reduction side path in a dual path compressor or expander or, alternatively, it may form a stand alone compressor. This arrangement is but one example of the way in which multiple elements with different characteristics can be combined. Another useful combination of elements is to provide for elements with upward acting and downward acting sliding band characteristics within the same frequency band.

A five element stack is shown in Figure 11 having a wideband attenuator element 68, a fixed band attenuator element 70, a sliding band element 72, a first spot frequency attenuator element 74 and a second spot frequency attenuator element 76. Each element receives the input signal. High pass filters 78 and 80, each having an 800 Hz corner frequency, form a part of the transfer functions of elements 70 and 72. Band pass filters 82 and 84, centered at 960 Hz (an audio interference frequency related to head rotation in quadruplex videotape recorders) and at 15.75 kHz (an audio interference frequency at the horizontal sweep frequency in 525 line television systems) form a part of the transfer functions of elements 74 and 76, respectively. Each element has its own independent control circuit 86, 88, 90, 92, and 94, essentially a rectifier and smoothing circuit including an amplifier which can have frequency weighted or wideband gain, as may be necessary. Preferably, each control circuit responds only to the output of its respective element; that is, the control signal is derived differentially using combining means 89, 91, 93, and 95 (note subtractive feed from bottom of each element). The control circuits control the variable resistance elements 96, 98, 100, 102, and 104, which, in practice may be FETs, the control voltage being applied applied to their gate input. The fixed band attenuator elements 68, 70, 74 and 76 have series resistors (106, 108, 110, and 112, respectively) to form a variable voltage divider in conjunction with the FETs. The sliding band element 72 has a series capacitor 114 to form a variable high pass filter in conjunction with the FET. In the manner of the preferred stack arrangement of Figure 4, element 68 is driven with respect to a reference, shown as ground and each of the other elements is driven through a buffer (unity gain buffers 116, 118, 120, and 122) with respect to the output of the element below it. Note that each filter, if included, also has its reference terminal connected to the output of the element below it. Each element has a threshold level such that when the signal level rises in the element and reaches that level a change in the variable element begins. Each element is assumed to provide a maximum of 10 dB of dynamic action. The overall output of the arrangement is from the output of the top element through a unity gain buffer 124.

In the quiescent state, the defining envelope of the arrangement of Figure 11 is a total dynamic action of 10 dB, wideband. As signal components appear, the characteristics of the various elements emerge in response to the frequencies and levels of the components. The sequence in which the characteristics of elements emerge in response to those signal components will depend on the operating frequency range of the respective elements and their relative threshold levels.

Consider first two examples. If a 200 Hz tone appears, starting at a sub-threshold level and gradually rising in level, the wideband element is eventually defeated, but the others remain fully effective. Thus the overall characteristic, shown in Figure 12A, is that of the 800 Hz fixed and sliding bands, which are coincident and partly hide the two spot frequency characteristics: the tip of the 960 Hz spot frequency characteristic and the sliding band characteristic, the high pass filter of the sliding band having moved upwardly in response to the 2 kHz tone. The sliding band characteristic hides the 15.75 kHz spot frequency characteristic.

In the arrangement of Figure 11, each element has its own control circuit which operates entirely in response to signals within the respective element as a result of the differential derivation of the control signals. Elements may have each have a totally independent control circuit, with or without differential derivation, and various cross connections are possible. The cross connections illustrated in Figures 13 and 14 are useful in permitting different thresholds to be set in the elements while retaining adequate control circuit gain in the sliding band element at frequency extremes. Another type of cross connection, referred to as "parking" is described below in connection with Figures 16A and 16B.

Referring to Figure 13, a fixed band element 126, a wideband attenuator, is shown stacked with a sliding band element 128, a sliding shelf acting upwardly in frequency and having a wideband quiescent characteristic: substitution of an inductive element instead of the capacitor would result in a sliding shelf acting downwardly in frequency. As a practical matter, a gyrator circuit would be used to simulate an inductor. The fixed band element has a series resistor 130 and a shunt FET 132, connected to a reference potential, which form a variable attenuator as the resistance of the source-drain path of the FET is controlled in response to a control voltage applied to its gate. The output of the fixed band element is applied through a buffer amplifier 134 to a rectifier and smoothing circuit 136. The sliding band element 128 includes parallel capacitor 138 and resistor 140 shunted by FET 142 which is connected to the output of the fixed band element, at the output of buffer 134. The output of the sliding band element is applied to a buffer amplifier 144, an optional combining means 150, and a rectifier and smoothing circuit 146 which provides a control voltage to the FET 142. The output of the combination of elements is taken from buffer amplifier 144.

As noted in connection with the arrangement of Figure 11, it is possible to use the output of one element in a stack to buck the output of the next element in generating the control voltage of that next element. This is shown not only in Figure 11 but also in Figure 13 where the output of the sliding band element from buffer 134 is subtracted from the output of the fixed band element from buffer 144 in the combining means 150. The bucking signal from the fixed band element can be treated so as to be more or less than the sliding band element output (by amplification or attenuation) and the sense of the bucking signal can be inverted so that as the fixed band element output increases the bucking signal decreases. These principles apply generally to successive active elements in any arrangement according to the invention.

A further alternative is shown in Figure 14 in which the bucking signal from the fixed band element is first conditioned by a filter 152 located in the control circuit loop for the fixed band element. Typically, the filter would be a high pass or low pass filter located within the operating band of the overall circuit. An unfiltered bucking signal, shown as a dashed line, may simultaneously be applied in the manner of the alternative version of Figure 13 (e.g., with optional combining means 150). The filtered bucking signal can be treated as mentioned in connection with the unfiltered bucking signal of Figure 13.

In a further control circuit coupling variation, it is possible to feed the rectified signal in the control circuit of one element into the DC portion of the control circuit of another element. For example, in Figure 13, the summing means 150 would be located between the output of rectifier and smoothing circuit 146 and the gate of FET 142, and the bucking signal from element 126 would be taken from the output of rectifier and smoothing circuit 136. This alternative is advantageous where the signal phase relations may not be correct for AC combining/bucking as in the method discussed above. However, a disadvantage is that it is not possible to apply selective filtering of the coupled component, as in Figure 14.

In some applications it may be possible to eliminate one control circuit and use a common control circuit for two elements as is shown in Figure 15, although the use of individual control circuits for each element is preferred to produce a more conformal effect. (The same reference numerals are used for common components in Figures 13 through 15.) In Figure 15, the output of the fixed band control circuit from the rectifier and smoothing circuit 136 is applied through a buffer 148 in order also to control the sliding band FET 142.

Figure 16A shows a modification of Figure 14 in which an additional feature referred to as "parking" is incorporated in the circuit. The concept of parking is based on the observation that it is often advantageous to pre-set or "park" a response characteristic at some level and/or frequency other than its quiescent level and/or frequency when the element is below its threshold. In the case of the fixed band element, the fixed band control circuit is parked at a level just below its threshold so that the circuit reacts more quickly in response to changes in the input signal. A detector 153 senses the fixed band control signal, develops an appropriate signal and applies it to summing means 155 where it is added to, or takes over from, the control signal to provide the fixed band parking signal. In operation, the parking signal level preferably is set just below the fixed band threshold. Above the threshold, the fixed band control circuit takes over to the exclusion of the parking detector. Thus, the detector 153 is required to sense the fixed band control signal

and to provide, only when the fixed band element is below its threshold, the correct signal level for parking the fixed band control circuit at the desired level. A detector providing these functions may be implemented in various ways: one suitable arrangement is to sense the fixed band control voltage with a "perfect diode" circuit. Such circuits are well known and employ an operational amplifier biased so that the circuit switches its output from positive to negative when the signal being sensed rises to the bias voltage. The perfect diode circuit output, suitably adjusted in level as may be necessary, acts as a maximum selector circuit which takes over from the fixed band control signal to provide the desired parking signal.

Figure 16B shows generally another arrangement of stacked fixed band and sliding band elements employing parking. In this case the sliding band element 129 has an inductor 139 (simulated by a gyrator circuit in practical embodiments) so that the sliding band circuit operates downwardly. In the arrangement of Figure 16B, parking is provided not only in the fixed band circuit but also in the sliding band circuit. As applied to the sliding band circuit, the concept of parking is based on the observation that as long as the fixed band umbrella characteristic is present, then it is possible to pre-slide the sliding band ("park" it) to the frequency region where the action is likely to be when a signal is applied. This improves response time and reduces modulation distortion. In its quiescent state the downwardly acting sliding band would normally be in its wideband state (in the absence of any parking circuit)--that is, the high frequency cutoff would be infinity, i.e., well into the low frequency stop-band if the circuit were employed in an arrangement such as described below in connection with Figure 19. When a signal appears the band slides down to attenuate the dominant signal, which with speech and music very likely will be in the range 100 Hz - 1 kHz. Therefore, it is useful to park the low frequency sliding band at a cutoff frequency of about 400 Hz, just into the low frequency pass-band of the Figure 19 arrangement. The parking influence is removed as soon as the signal rises above the fixed band threshold as this marks the imminent reduction or disappearance of the fixed band umbrella. Detector 153 senses the fixed band control signal in relation to a reference voltage and develops the parking signal which is applied to the sliding band FET 142 via combining means/maximum selector 151. After the signal has risen above the fixed band threshold, if the applied frequency is lower than 400 Hz, the sliding band control circuit takes over and slides the band down further. If the frequency is above 400 Hz, the sliding band control circuit still takes over but allows the cutoff frequency to slide up. The detector 153 is required to sense the fixed band control signal and to provide, only when the fixed band element is below its threshold, the correct signal level for parking the sliding band circuit at the desired frequency. The detector described in connection with the arrangement of Figure 16A is also suitable for the arrangement of Figure 16B.

While sliding band parking arrangements can be used in connection with high frequency sliding band circuits as well, the advantages are less because the likely applied signal frequencies are in the high frequency stop band frequency range (referring to arrangements such as in Figures 19 and 20), which is already the right way around, in contrast with the low frequency situation.

Figure 16B shows an additional feature, related to the parking circuits described above, for reducing noise generated by the gyrator simulated inductance 139. A shunt shown as switch 157 is connected across the inductor 139. The switch is controlled by the signal from detector 153. Below the thresholds of the fixed band and sliding band circuits, as sensed by detector 153, there is no need for the sliding band circuit 129 and the inductor 139 to be active. Thus, the gyrator simulated inductance is shorted out by the shunt when the circuits are below their threshold. In a practical circuit, the shunt can be implemented using an FET, the signal from detector 153 being applied to the FET gate. Alternatively, the shunt could be located across any input point to an appropriate output point, the two points having the same signal levels. The use of a shunt to eliminate noise in an active element could be used in any bi-linear or uni-linear compressor or expander circuits because in such circuits there is a range of input signal levels in which the circuit provides no dynamic action. Such a shunt is not necessarily related to parking circuits. In the arrangement of Figure 16B the parking signal derived from detector 153 merely happens to be suitable for control of the shunt.

The parking and the gyrator shunt features of the embodiments of Figures 16A and 16B need not be employed in the context of the Figure 14 arrangement. For example, the use of a filter 152 is not essential. The parking feature could be applied to a general arrangement such as that of Figure 11.

The cross coupling arrangements of Figures 13 through 16B thus provide considerable flexibility for the circuit designer in determining the manner in which the thresholds of the two elements will interact.

In principle, where the stacked elements each have separate and independent control circuits with full differential control, such as in the arrangement of Figure 13, the order of the elements in the stack is immaterial: the same results should be obtained with the elements arranged in any order. However, where there is no cross coupling in the control circuitry, or a cross coupling that provides only a partial differential control (as in Figure 14 when the only bucking signal is from filter 152), the order in which the elements are stacked may affect the overall result.

In order to understand the operation of a two element stack made up of fixed band and sliding band elements (such as in Figures 13 through 16), assume that a fixed single pole high pass filter with a corner frequency of 800 Hz, for example, is placed in series to the inputs to the two elements and that the effective threshold of the fixed band element is -62 dB and that of the sliding band element is -66 dB. Each element is assumed to provide a maximum of 10 dB of attenuation. If, for example, a 12 kHz signal is applied at a level of -66 dB, the corner frequency of the sliding band characteristic begins to slide but nothing happens to the overall characteristic envelope because the fixed band is still inactive and supports the envelope. The altered sliding band characteristic is hidden. This situation, shown in Figure 17A, which has the same envelope as the quiescent condition, but showing the altered sliding band characteristic in phantom, continues until the signal reaches the -62 dB threshold of the fixed band element. As the signal rises above the fixed band threshold (by a few dB), the fixed band begins to attenuate revealing the top of the sliding band characteristic, which continues to slide upward as the signal level rises. This situation is shown in Figure 17B. As the signal level increases several more dB above the fixed band threshold, the fixed band continues to attenuate and the sliding band continues to move upward as shown in Figure 17C.

As discussed in US-PS Re 28,426, in order to obtain full attenuation of the sliding band FET, it is necessary to have a sufficiently high gain in the control signal loop at high levels. The solution set forth in that patent is to provide high frequency weighted amplification in the loop. However, a drawback to that approach is that the loop gain is too high at lower levels (thus causing the band to slide farther than necessary) and it is also difficult to maintain a sufficiently high sliding band threshold.

By employing the cross coupling techniques described in connection with Figures 13 and 14 it is possible to achieve the desired sliding band threshold with the desired loop gains at high and low signal level conditions. If the arrangement of Figure 13 with the optional combining means 150 is used (i.e., full bucking effect), then below the fixed band threshold there would be no control signal drive at all to the sliding band FET and under very high level conditions when the fixed band is defeated (e.g., full attenuation), no further bucking is provided and full gain is provided in the sliding band control signal loop.

In practice, the full bucking approach provides more bucking effect than is necessary and improved performance is achieved if the bucking action is allowed to be less. For example, the effective high frequency gain of the sliding band control signal loop may be reduced by about 5 dB as long as the fixed band is not attenuated and then as it attenuates the bucking action is decreased until it reaches 0 dB when the fixed band is fully attenuated. In critical applications, where a high sliding band threshold is desired, the arrangement of Figure 14 allows the use of both wideband and frequency selective bucking action. In the case of a high frequency sliding band element, the frequency selective bucking signal is derived using a high pass filter (filter 152, Figure 14) since the bucking action is most desirable at the high frequency extreme, where the equalized control amplifier provides the highest gain. In one practical embodiment described further below a cutoff frequency of 1.6 kHz is employed. For a low frequency sliding band element, the requirement is to provide high loop gain at high level low frequencies. In that case, the filter 152 in the arrangement of Figure 14 is a low pass filter. In one practical embodiment described further below a cutoff frequency of 400 Hz is employed.

As mentioned above, a very useful arrangement is to provide a compressor or expander configured in the nature of bandsplitting high frequency and low frequency stages, each stage made up of sliding band and fixed band elements. The parallel stages can operate as stand alone dynamic range modification devices or as side paths in Type I or Type II dual path arrangements such as described in US-PS 3,846,719; US-PS 3,903,485; US-PS Re 28,426; and US-PS 4,490,691. Preferably, single pole filters with the same corner frequency in each are employed so that the combined quiescent response of the two stages is flat. An advantageous corner frequency for the high and low frequency stages is about 800 Hz for an audio system. With the gently sloping 6 dB/octave slope of the filters, if the high frequency stage has a corner frequency of 800 Hz, the stage provides significant action as low as 100 or 200 Hz. If the low frequency stage also has a corner frequency of 800 Hz, it provides significant action as high as 3 kHz to 6 kHz. Thus, for signals lying in the range of, say 100 Hz to 6 kHz, which is the band having most of the energy in typical music, both stages are effective and provide a signal tracking effect. For example, as explained further below, for a single dominant signal in that range, the overall response is that of two sliding bands, above and below the dominant signal. For more than one dominant signal the resulting characteristic is a fixed band response between the most upper and most lower dominant signals and sliding band responses above and below the most upper and most lower dominant signals, respectively.

In Figure 18, an embodiment of a high frequency stage is shown. A single pole high pass filter 154 with a cutoff frequency of 800 Hz is located in its input. In practice the filter includes an operational amplifier so as to buffer the filter from the following circuitry. The high pass filtered input signal is applied to a fixed band element 156 and to a sliding band element 158. The fixed band element includes an input resistor

160, a shunt FET 162 operated as a variable resistor, and a control circuit 164 that generates a DC control signal which is applied to the gate of the FET 162. The FET resistance drops as the DC control signal level increases. Fixed band control circuit 164 includes in its loop a buffer amplifier 166, a single pole high pass filter 168 having about a 400 Hz cutoff frequency, a rectifier 172 (preferably full-wave), and a smoothing circuit 174 (which also is employed to adjust the attack and release time constants of the control loop). The sliding band element 158 includes parallel input resistor 178 and capacitor 180 which are shunted by FET 182. The sliding band element has a control circuit 184 which includes a buffer amplifier 186, a summing means 188, a high frequency emphasis circuit 190, a rectifier 192 (preferably full-wave), and a smoothing circuit 194. The summing circuit also receives as inputs bucking signals taken before and after a high pass filter 170 having about a 1.6 kHz cutoff frequency. The bucking signals are suitably buffered to avoid undesired interaction between the control circuits. The input to filter 170 is taken between filter 168 and rectifier 172. The output of the fixed band element 156 is applied to the sliding band element through buffer 198. The overall output is taken from buffer 186 of the sliding band element 158.

In operation, the stacked fixed band and sliding band elements operate generally in the manner described above. The coupling of bucking signal components from the fixed band element to the control circuit of the sliding band element are for the reasons set forth in connection with the description of Figure 14, above. A corner frequency of about 1.6 kHz was chosen for filter 170 because differential control of the sliding band is needed only at high frequencies. Filter 168 is employed in order to reduce the sensitivity of the fixed band circuit to very low frequency components.

In Figure 19, an embodiment of a low frequency stage is shown. The input signal is applied to a fixed band element 200 and to a sliding band element 202. The fixed band element includes an input resistor 204, a shunt FET 206 operated as a variable resistor, and a control circuit 208 that generates a DC control signal which is applied to the gate of the FET 206. The FET resistance drops as the DC control signal level increases. Control circuit 208 includes in its loop a buffer amplifier 210, a first single pole low pass filter 212 having a corner frequency of about 800 Hz, a second low pass filter 214 having a corner frequency of about 1.6 kHz, a rectifier 218 (preferably full-wave), and a smoothing circuit 220 (which also is employed to adjust the attack and release time constants of the control loop).

The sliding band element 202 includes parallel input resistor 224 and inductor 226 which are shunted by FET 228. In practice, the inductor 226 is simulated by a gyrator circuit which includes operational amplifiers (such circuits are well known and is not shown). The sliding band element has a control circuit 230 which includes a buffer amplifier 232, a single pole low pass filter 234 having a corner frequency of about 800 Hz, a summing circuit 236, a low frequency emphasis circuit 238, a rectifier 240 (preferably full-wave), and a smoothing circuit 242. Low pass filter 234 is preferably located in the position shown in order to assist in suppressing undesired noise generated in the low frequency stage. Alternatively, the filter may be located in the input to the arrangement as is done in the high frequency circuit of Figure 18. The summing circuit 236 also receives as inputs bucking signals taken before and after a low pass filter 216 having a corner frequency of about 400 Hz. The bucking signals are suitably buffered to avoid undesired interaction between the control circuits. The input to filter 216 is taken between filter 214 and rectifier 218. The output of the fixed band element 200 is applied to the sliding band element through buffer 247. The overall output is taken from filter 234 of the sliding band element 202.

In operation, the low frequency stacked fixed band and sliding band elements operate generally in the manner described above except that the sliding band element operates downwardly in frequency. One difference in the low frequency stage, as noted above, is that the band defining filter is located in the stage output rather than in the input as in the high frequency stage. The coupling of bucking signal components from the fixed band element to the control circuit of the sliding band element are for the reasons set forth in connection with the description of Figure 14, above. A corner frequency of about 400 Hz was chosen for filter 216 because differential control of the sliding band is needed only at low frequencies. Filters 212 and 214 are employed in order to reduce the sensitivity of the fixed band circuit to very high frequency components.

High frequency and low frequency stages of the type set forth in Figures 18 and 19 can be used as building blocks in creating compressors, expanders and noise reduction companders. For example, high frequency and low frequency stages of the type described in Figures 18 and 19 may be employed as side paths in dual path arrangements in the manner shown in Figures 20 and 21.

In Figure 20 a Type I dual path arrangement (of the type generally described in US-PS 3,846,719) is shown having a compressor 248 in which the input signal is applied to the high frequency stage 250, to the low frequency stage 252 and to the main path 254. The outputs of stages 250 and 252 are summed in summing means 256 and then summed with the main path signal components in summing means 258 to provide the compressor output for application to a transmission channel. The side path signal components

thus boost the main path signal components causing compressor action. The transmission channel output is applied to the expander 260, configured in a complementary manner to the compressor 248, which has an input summing means 262 which receives the transmission channel output and subtracts the sum of the high frequency stage 250 and low frequency stage 252 outputs, which are added in summing means 264.

6 The side path signal components thus buck the main path signal components causing expander action.

In Figure 21 a Type II dual path arrangement (of the type generally described in US-PS 3,903,485) is shown having a compressor 268 which has an input summing means 270 receiving the input signal and the sum of the high frequency stage 250 and low frequency stage 252 outputs, which are combined in summing means 272. The summing means 272 has its output applied to the main path 274 which provides
10 the compressor output to the transmission channel and the input to the stages 250 and 252 of the compressor. The side path signal components thus boost the main path signal components causing compressor action. The transmission channel output is applied to the expander 276, configured in a complementary manner to the compressor 258. The input signal is applied to the high frequency stage 250, to the low frequency stage 252 and to the main path 278. The outputs of stages 250 and 252 are summed
15 in summing means 280 and then subtracted from the main path signal components in summing means 282 to provide the expander output. The side path signal components thus buck the main path signal components causing expander action.

In Figures 20 and 21 the main path of each compressor and expander is linear with respect to dynamic range and the level of the sum of the side path high frequency and low frequency stages is less than the
20 maximum level of the main path. The transmission channel in those Figures and in Figures 22 and 23, below, may include any type of storage or transmission medium and may also include means for converting or encoding the analog signal components from the compressor into a different form (digital, for example), the storage or transmission of the encoded signals, and means for re-converting or decoding the encoded signals back into analog signal components.

In arrangements such as in Figures 20 and 21 where only one high frequency stage and one low frequency stage is used in each compressor and expander it is practical to provide a maximum of about 10 to 12 dB of noise reduction without reaching excessive maximum compression or expansion ratios. Although the arrangements of Figures 20 and 21 will be adequate in certain applications, it is useful to employ the teachings of US-PS 4,490,691 to achieve a greater amount of overall noise reduction without
25 placing an undue burden on any one stage or creating excessive compression or expansion ratios.

In Figure 22 one possible arrangement is shown in which there are three series Type I dual path stages in the compressor and three complementary stages in the expander. The threshold levels of the series bi-linear circuits are staggered, employing the staggering aspects of US-PS 4,490,691. Alternatively, a Type II configuration could be employed. The embodiment of Figure 22 also employs the spectral skewing and
30 antisaturation aspects of US-PS 4,490,691, although these aspects are not essential to a multi-stage arrangement employing high frequency and low frequency stages.

The compressor portion of the system of Figure 22 has three stages: a high level stage 284, which has the highest threshold level; a mid-level stage 286; and a low level stage 288, which has the lowest threshold level. As discussed in US-PS 4,490,691 this is the preferred order of arrangement of staggered stages,
40 although the reverse order is possible. The expander portion of the system of Figure 22 also has three stages arranged complementary to the compressor: the low level stage 290, the mid-level stage 292, and the high level stage 294. Each high level and mid-level stage has both a high frequency stage 250 and a low frequency stage 252. The low level stage has only a high frequency stage 250 and no low frequency stage. Each high frequency stage 250 and each low frequency stage 252 is of the type described in
45 connection with Figures 18 and 19. In practical circuits there may be some differences between or among high frequency and low frequency stages depending on whether it is located in the high level, mid-level or low level stage.

If each compressor stage (284, 286, 288) and each expander stage (290, 292, 294) has, for example, 8 dB of compression or expansion, respectively, then the overall compander system will provide 24 dB of
50 noise reduction in the high frequency band (above 800 Hz, if the high frequency stages have an 800 Hz cutoff frequency) and 16 dB of noise reduction in the low frequency band (below 800 Hz, if the low frequency stages have an 800 Hz cutoff frequency). Such an arrangement is useful, for example, in a high quality audio noise reduction system of the type used in professional applications.

The cascading of staggered stages as in the embodiment of Figure 22 is useful not only to increase the
55 amount of compression and expansion but also to create a quasi-multi-pole effect (with respect to the band of the high frequency stages) and a quasi-two-pole effect (with respect to the band of the low frequency stages). Thus, while each circuit itself has only a single pole band determining filter, in order to allow for broad frequency overlap of the circuits and a flat overall frequency response when the bands are combined,

the overall effect when two or three circuits are cascaded is to provide a steepness compounding or stage ganging effect: in effect, a two or three pole filter. The additional sharpness of a two or three pole filter is desirable in that it allows the system to track dominant signal components more effectively. That is, with steep filter slopes, frequencies adjacent to a dominant signal component are affected as little as possible by dynamic action. The quasi-two-pole and three-pole filter effect brings about that result while retaining the advantages of single pole filters in each individual circuit, namely that the outputs of the high and low frequency circuits combine perfectly in both amplitude and phase. The overall result is that the filter action becomes more complex as signals are applied by virtue of the multi-stage steepness compounding effect taken in combination with the quasi-two-pole filter effect in the noise reduction side paths of each circuit.

The latter effect results from the use of a single pole fixed filter in series with a variable shelf characteristic in the sliding band portion of the circuits which has a single pole characteristic in its quiescent condition but which tends toward a two-pole characteristic in the presence of a dominant signal or signals. The intervention of the fixed bands, via action substitution, further tailors the response.

The input to the compressor portion of the system is applied to low frequency and high frequency spectral skewing networks shown as block 296. In a practical embodiment there are two networks in series: a two section low frequency network made up of a 50 Hz high pass single pole shelving section and a 20 Hz high pass single pole filter section; and a 12 kHz Butterworth (maximally flat) low pass filter network. Both networks may be implemented using well known operational amplifier active filter techniques. Complementary de-skewing networks are located in block 304 at the output of the expander.

The main path of each stage 284, 286, and 288 in the compressor portion includes low frequency and high frequency antisaturation networks 298, 300, and 302, respectively. Complementary antisaturation networks 304, 306, and 308 are located in the main path of each stage 290, 292, and 294, respectively, in the expander portion. As discussed in US-PS 4,490,691, it is possible to locate such networks in the main path of only one stage in the compressor and in the complementary location in one stage in the expander portion of a series of cascaded staggered stages.

The Type I stages of Figure 22 also include summing means 310, 312, 314 and 316 that combine the outputs of the high frequency and low frequency circuits in stages 284, 286, 292, and 294, respectively. The stages each include summing means 318, 320, 322, 324, 326, and 328 in the main paths which couple to the main path the side path output, in the case of stages 288 and 290, or the outputs of the side paths, in the case of the other stages.

The arrangement of Figure 22 is only one example of compressor, expander or compander systems that can be constructed using the high frequency circuits and/or the low frequency circuits of the type described in connection with Figures 18 and 19. Other possibilities include, but are not limited to, (1) a single high frequency circuit in the compressor and expander; (2) one high frequency circuit and one low frequency circuit each in the compressor and expander (as in the examples of Figures 20 and 21); and (3) two high frequency circuits and one low frequency circuit each in the compressor and expander (e.g., an arrangement such as in Figure 22, but omitting the high level stages 284 and 294). Alternative (1) would provide a degree of compatibility with widely used B-type consumer noise reduction systems licensed by Dolby Laboratories and alternative (3) would provide a degree of compatibility with the widely used A-type professional noise reduction systems manufactured by Dolby Laboratories. Other configurations are possible depending on the goals of the system designer. For example, a fourth stage could be added to the compressor and expander portions of the example of Figure 22 and/or the low level stage of that example could additionally include a low frequency circuit.

In a practical embodiment of a system generally in accordance with the example of Figure 22 it is possible to employ only one set of high level, mid-level and low level stages by using a unique circuit and switching arrangement. Figure 23 is a block diagram of such a practical embodiment. The circuit and switching arrangement is applicable to all types of compander systems having series dual path stages in which simultaneous encoding and decoding is not required. A single pole double throw switch provides switching of the system from its compression mode to its expansion mode. In order to provide this simplicity in switching, the arrangement employs a supplemental main path in each stage except the last.

Referring to Figure 23, the input to the system is through a bandpass filter 330 which is intended to remove subsonic and supersonic signals (for example, the filter may be comprised of a two pole high pass filter with about a 10 Hz cutoff frequency and a two pole low pass filter with about a 50 kHz cutoff frequency). The filtered input signal passes through first and second inverting buffer amplifiers 332 and 334 (configured as summing operational amplifiers with appropriate resistors 332a, 332b, 332c, 334a, 334b, and 334c) to the system output. The main path of the system is between amplifiers 332 and 334. At a point between buffers 332 and 334 the input to three paths is derived; the three paths include a first path 336 including a low frequency and high frequency spectral skewing network 338 feeding three stages (340, 342,

344), the first two (340, 342) having supplemental main paths (346, 348), a second path 350 which includes a low frequency and high frequency antisaturation network 352 and generates an antisaturation bucking signal, and a gain trimming path 354. Network 338 has the same characteristics described in connection with network 296 of Figure 22. The effect of network 352 is to create similar characteristics as networks 298, 300, and 302 of Figure 22, but further includes amplification means as may be necessary to generate a bucking signal of the proper level and polarity. Signal components from the three paths are combined in summing means 356, the output of which is applied to the switching arrangement 358. That is, network 352 has a low frequency and a high frequency boost characteristic so that when the signal is combined in a bucking sense the overall low frequency and high frequency antisaturation characteristics are obtained.

The first stage 340 is preferably the high level stage which includes a high frequency circuit 250 and a low frequency circuit 252, as in Figure 22. The input to stage 340 is fed to circuits 250, 252 and the secondary main path 346. The output of each (e.g., 250, 252, 346) are combined in summing means 364. The second stage 342, the mid-level stage, receives the output of summing means 364 and applies that to further high frequency and low frequency stages, 250 and 252, respectively, and to secondary main path 348. The output of circuits 250, 252 and path 348 are combined in summing means 366 which applies its output to the low level stage 344 which consists of a high frequency circuit 250. The noise reduction outputs of the five circuits in stages 340, 342 and 344 are applied to combining means 356 along with the anti-saturation bucking signal from block 352 and the level trimming path 354. The level trimming is in the order of a fraction of a decibel.

The switching arrangement includes a single pole single throw noise reduction on/off switch 360 which grounds the output from summing means 356 in its closed, noise reduction "off", position and which is open in the noise reduction "on" position. The summing means 356 output is further applied to the common position of a single pole double throw switch 362. In the "compress" position, the summing means output is applied to a point between buffers 332 and 334. In the "expand" position, the summing means output is applied to the input of buffer 332.

In operation, closure of switch 360 to the noise reduction "off" position prevents the output of summing means 356 from being coupled to buffer 334 because it grounds one of its two summing inputs. As to the compress/expand mode switch 362, in the "compress" position the overall arrangement is essentially equivalent to that of a Type I compressor as shown in a portion of Figure 22. That is, the input signal via buffer 332 is applied to the three paths 336, 350 and 354 and their sum is applied to the output via switch 362 and buffer 332. In the expand position of switch 362, the output of combining means 356 is fed back and is of opposite polarity to the input to the three paths, thus replicating the Type I expander configuration of Figure 22. In either switch position, the output of combiner 356 is added to the input signal by amplifier 332 or amplifier 334. The same principles can be applied to a Type II compander system.

In addition to simplifying compressor/expander switching, the arrangement of Figure 23 only requires a single spectral skewing network and a single antisaturation network. The use of a single spectral skewing network in the arrangement of Figure 23 does result in one minor difference from the configuration of Figure 22, namely that in Figure 23 the network 338 affects only the side paths and does not affect the primary main path whereas in Figure 22 the network 296 affects both the main and side paths. However, as set forth in US-PS 4,490,691, both locations are acceptable alternatives.

In a further practical embodiment of a system generally in accordance with the example of Figure 22 it is also possible to employ only one set of high level, mid-level and low level stages by using a unique circuit and switching arrangement. Figure 24 is a block diagram of such a practical embodiment. The circuit and switching arrangement is applicable to all types of compander systems having series dual path stages in which simultaneous encoding and decoding is not required. A single pole double throw switch provides switching of the system from its compression mode to its expansion mode. A further switch provides a noise reduction off position. Even greater simplicity is provided than in the embodiment of Figure 23 in that no supplemental main paths are required. The amount of antisaturation required may be provided by only one low frequency and one high frequency antisaturation network as shown. The arrangement of Figure 24 utilizes the encoder portion of Figure 22 in parallel with an inverting amplifier such that the parallel combination when summed provides a derived noise reduction side path signal, e.g., the parallel combination effectively subtracts the main path signal from the main path plus noise reduction signal to provide the noise reduction signal by itself. By deriving the noise reduction signal in isolation, the overall system is easily switched from an encoding mode to a decoding mode as explained in connection with Figures 24 and 25.

Referring to Figure 24, in which the same reference numerals are used as for like elements in Figures 22 and 23, the input to the system is through bandpass filter 330. The filtered input signal passes through first and second inverting buffer amplifiers 362 and 364 (configured as summing operational amplifiers with

appropriate resistors 362a, 332b, 364a, and 364b) to the system output. At a point between buffers 362 and 364 the input to the encoder (compressor) portion of the system of Figure 22 in parallel with the inverting amplifier 370 is derived. The main path of the system is provided between amplifier 362 and amplifier 364. The encoder output is applied to a summing resistor 368. The inverting amplifier path includes inverting operational amplifier 370, an input resistor 370a, feedback resistor 370b, and summing resistor 372. Resistors 370a and 370b are selected to provide unity gain. Summing resistors 368 and 372 have the same value so that when the encoder and inverting amplifier paths are summed in either amplifier 362 or 364 the resulting signal is the derived noise reduction signal.

Compress/expand switching is provided by a single pole two position switch 374 that has its common point connected to the junction of resistors 368 and 372. In the "compress" position, the summing resistors 368 and 372 are connected to the input of summing amplifier 364. The amplifier therefore sums the noise reduction signal with the main path signal to provide compression. In the "expand" position, the summing resistors 368 and 372 are connected to the input of summing amplifier 362. The amplifier therefor subtracts the noise reduction signal from the main path signal to provide expansion. In the "off" position of switch 375, the noise reduction signal is not connected and the output is simply the input.

The operation of the switching arrangement of Figure 24 may be better understood by reference to Figure 25 which shows the prior art switching arrangement for a single stage type I dual path compander system as described in US-PS Re 28,426. The input signal is applied to a first summing means 374. The main path 376 extends from summing means 374 to a second summing means 378. The noise reduction side path 380 (heretofore a single sub-circuit) derives its input from the main path and has its output switchable by switch 380 so as to sum with the main path in summing means 378 for compression and to subtract from or buck the main path by inversion in inverter 384 and summation in summing means 374. Switch 383 opens the NR side path for the off condition. The arrangement of Figure 24 is based on the recognition that it is possible to simply derive a noise reduction side path signal representative of the overall noise reduction signal in a complex multi-stage system.

The arrangement of Figure 24 can be modified by employing the decoder portion of Figure 22 in place of the encoder portion. In addition, these principles are equally applicable to type II arrangements.

It will be noted that there is a further advantage of the arrangement of Figure 24 over the arrangement of Figure 23: the effect of the spectral skewing network in the arrangement of Figure 24 is that same as in the configuration of Figure 22; it affects both the main paths and the side paths. In this way, the spectral skewing network also provides an antisaturation effect at the extreme ends of the audio spectrum. This effect is over and above that provided by the antisaturation networks themselves.

Figure 26 is a representation of the quiescent characteristics of the five circuits and the overall arrangement for a system configured such as the compressor portion of Figure 22 or the systems of Figure 23 and 24 operated in their compressor mode, in each case without the respective spectral skewing and anti-saturation networks. Each circuit 250 and 252 provides a maximum of 8 dB of compression for low level signal conditions below threshold. Thus, the curves show the additive effect of the two low frequency circuits and of the three high frequency circuits, providing an overall compression of 24 dB at high frequencies dropping to 16 dB at low frequencies. The curves also illustrate the substantial overlap of the 6 dB/octave skirts of the single pole filters. Reference should also be made to Figure 7B described above, which illustrates the low level response of the same arrangement in the presence of dominant signals.

Claims

1. A circuit for modifying the dynamic range of an input signal in a frequency band, comprising a plurality of circuit elements (14-20; 30-36; 38-44; 58-62, 66), each circuit element having a voltage transfer characteristic represented by $t_1(s)$, $t_2(s)$, . . . $t_n(s)$, where n is the number of said circuit elements, said transfer characteristic being frequency-dependent, or dynamic range modifying, or both frequency-dependent and dynamic range modifying, at least one of said circuit elements having at least a dynamic range modifying transfer characteristic, said circuit elements having transfer characteristics that operate at least partially within the same frequency and level region, means (22-28; 46-56; 64) for interconnecting the circuit elements such that when an input signal represented by the voltage V_{in} is applied, the output voltage V_{out} is represented by

appropriate resistors 362a, 332b, 364a, and 364b) to the system output. At a point between buffers 362 and 364 the input to the encoder (compressor) portion of the system of Figure 22 in parallel with the inverting amplifier 370 is derived. The main path of the system is provided between amplifier 362 and amplifier 364. The encoder output is applied to a summing resistor 368. The inverting amplifier path includes inverting operational amplifier 370, an input resistor 370a, feedback resistor 370b, and summing resistor 372. Resistors 370a and 370b are selected to provide unity gain. Summing resistors 368 and 372 have the same value so that when the encoder and inverting amplifier paths are summed in either amplifier 362 or 364 the resulting signal is the derived noise reduction signal.

Compress/expand switching is provided by a single pole two position switch 374 that has its common point connected to the junction of resistors 368 and 372. In the "compress" position, the summing resistors 368 and 372 are connected to the input of summing amplifier 364. The amplifier therefore sums the noise reduction signal with the main path signal to provide compression. In the "expand" position, the summing resistors 368 and 372 are connected to the input of summing amplifier 362. The amplifier therefor subtracts the noise reduction signal from the main path signal to provide expansion. In the "off" position of switch 375, the noise reduction signal is not connected and the output is simply the input.

The operation of the switching arrangement of Figure 24 may be better understood by reference to Figure 25 which shows the prior art switching arrangement for a single stage type I dual path compander system as described in US-PS Re 28,426. The input signal is applied to a first summing means 374. The main path 376 extends from summing means 374 to a second summing means 378. The noise reduction side path 380 (heretofore a single sub-circuit) derives its input from the main path and has its output switchable by switch 380 so as to sum with the main path in summing means 378 for compression and to subtract from or buck the main path by inversion in inverter 384 and summation in summing means 374. Switch 383 opens the NR side path for the off condition. The arrangement of Figure 24 is based on the recognition that it is possible to simply derive a noise reduction side path signal representative of the overall noise reduction signal in a complex multi-stage system.

The arrangement of Figure 24 can be modified by employing the decoder portion of Figure 22 in place of the encoder portion. In addition, these principles are equally applicable to type II arrangements.

It will be noted that there is a further advantage of the arrangement of Figure 24 over the arrangement of Figure 23: the effect of the spectral skewing network in the arrangement of Figure 24 is that same as in the configuration of Figure 22; it affects both the main paths and the side paths. In this way, the spectral skewing network also provides an antisaturation effect at the extreme ends of the audio spectrum. This effect is over and above that provided by the antisaturation networks themselves.

Figure 26 is a representation of the quiescent characteristics of the five circuits and the overall arrangement for a system configured such as the compressor portion of Figure 22 or the systems of Figure 23 and 24 operated in their compressor mode, in each case without the respective spectral skewing and anti-saturation networks. Each circuit 250 and 252 provides a maximum of 8 dB of compression for low level signal conditions below threshold. Thus, the curves show the additive effect of the two low frequency circuits and of the three high frequency circuits, providing an overall compression of 24 dB at high frequencies dropping to 16 dB at low frequencies. The curves also illustrate the substantial overlap of the 6 dB/octave skirts of the single pole filters. Reference should also be made to Figure 7B described above, which illustrates the low level response of the same arrangement in the presence of dominant signals.

Claims

1. A circuit for modifying the dynamic range of an input signal in a frequency band, comprising a plurality of circuit elements (14-20; 30-36; 38-44; 58-62, 66), each circuit element having a voltage transfer characteristic represented by $t_1(s)$, $t_2(s)$, . . . $t_n(s)$, where n is the number of said circuit elements, said transfer characteristic being frequency-dependent, or dynamic range modifying, or both frequency-dependent and dynamic range modifying, at least one of said circuit elements having at least a dynamic range modifying transfer characteristic, said circuit elements having transfer characteristics that operate at least partially within the same frequency and level region, means (22-28; 46-56; 64) for interconnecting the circuit elements such that when an input signal represented by the voltage V_{in} is applied, the output voltage V_{out} is represented by



$$V_{out} = V_{in} \left[t_1(s) + \sum_{r=2}^n t_r(s) \left(\prod_{j=1}^{r-1} (1 - t_j(s)) \right) \right].$$

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2. A circuit according to claim 1 wherein said means for interconnecting the circuit elements interconnects said circuit elements (14-20) such that the output of the circuit is effectively the sum of the individual outputs of said circuit elements (22) and the effective input to each circuit element is a differential signal derived from said input signal and the individual outputs of one or more of the other circuit elements or a reference potential.
3. A circuit according to claim 1 wherein said circuit elements (30-36) are each represented as a three-terminal network wherein each first terminal and each second terminal are input terminals, and each third terminal is an output terminal, the transfer characteristic from the second terminal to the third terminal being the complement of the transfer characteristic from the first terminal to the third terminal, the input signal being coupled to the first terminal of each of said circuit elements, said means for interconnecting the circuit elements interconnecting the circuit elements in an arrangement wherein the second terminal of a first circuit element (30) is coupled to a reference potential, and the third terminal of each circuit element (30-34) is coupled to the second terminal of a successive circuit element (32-36), and the output is derived from the third terminal of the last circuit element (36).
4. A circuit according to claim 3 wherein said means for interconnecting the circuit elements interconnects the circuit elements in an arrangement that includes at least one branch, such that the third terminal of a circuit element (58) is coupled to the second terminal of two or more circuit elements (60, 62) which in turn have the output derived from their third terminal or have their third terminal coupled to the second terminal of one or more circuit elements (66).
5. A circuit according to any of claims 1 through 4 wherein
the circuit is for compressing the dynamic range of the input signal,
there are at least two circuit elements (126, 128) having a dynamic range modifying transfer characteristic, the dynamic range modifying transfer characteristic being a dynamic range reducing characteristic, and
the circuit provides characteristic action substitution,
such that transmission of very low level components of the input signal through the circuit tends to be maximized.
6. A circuit according to any of claims 1 through 4 wherein
the circuit is for expanding the dynamic range of the input signal,
there are at least two circuit elements (126, 128) having a dynamic range modifying transfer characteristic, the dynamic range modifying transfer characteristic being a dynamic range increasing characteristic, and
the circuit providing characteristic action substitution,
such that transmission of very low level components of the input signal through the circuit tends to be minimized.
7. A circuit according to any of claims 1 through 6 wherein there are at least two circuit elements (126, 128) having a dynamic range modifying transfer characteristic, the dynamic range modifying transfer characteristic responding to the level and/or spectral content of signal components in the respective circuit element above a threshold.
8. A circuit according to claim 7 wherein there are two circuit elements (126, 128), each having a dynamic range modifying transfer characteristic one of said circuit elements (126) being a fixed band circuit element, modifying dynamic range within a fixed frequency band and the other of said circuit elements (128) being a sliding band circuit element, modifying dynamic range within a sliding frequency band.



9. A circuit according to claim 8 wherein the sliding band circuit element (129) includes a gyrator-simulated component (139) which inherently generates undesired noise, the circuit further comprising means (157) for making the gyrator-simulated component ineffective during signal conditions when the gyrator-simulated component does not affect the dynamic action of the circuit.
10. A circuit according to claims 8 or 9 wherein the fixed-band circuit element (126) includes a control circuit (136) and the circuit further comprises means (153, 155) for maintaining said control circuit at a level just below its threshold level when the level and spectral content of the input signal are such that the fixed band circuit element is below its threshold.
11. A circuit according to claims 8, 9, or 10 wherein the dynamic range modifying transfer characteristic of the sliding band circuit element has a quiescent condition, the circuit further comprising means (151, 153) for moving the said characteristic away from its quiescent condition when the level and spectral content of the input signal are such that the fixed band circuit element is below its threshold.
12. A circuit according to any of claims 7 through 11 further comprising means (150) for coupling signal components between or among said circuit elements having a dynamic range modifying transfer characteristic to affect the threshold level of one or more of said circuit elements.
13. A circuit according to claim 12 wherein said means for coupling signal components includes filter means (152) for restricting the frequency band of the coupled signal components.
14. A circuit according to claims 12 or 13 wherein said circuit elements having a dynamic range modifying transfer characteristic include means for deriving a control signal (136, 146), said means for coupling signal components (150) including means for differentially deriving the control signal within a circuit element so that the control signal is substantially unaffected by the signal output(s) of another circuit element or other circuit elements.
15. A circuit according to any of claims 1 through 14 wherein the circuit is a dual-path circuit in which said plurality of circuit elements and said means for interconnecting the circuit elements constitute a further path (250), the circuit further comprising a main path (254) which is linear with respect to dynamic range, a combining circuit (258) in the main path, said further path having its input coupled to the input of the main path (254) or to the output of the main path (263) and its output coupled to the combining circuit (258), the further path providing a signal which, at least in a part of the frequency band, boosts or bucks the main path signal by way of the combining circuit, but which is so limited that, in the upper part of the input dynamic range, the further path signal is limited to a value smaller than the main path signal.
16. A circuit according to claim 15 wherein the further path includes a plurality of series circuits (284-288), at least one of said series circuits being a dual-path circuit, the circuit further comprising inverting means (370, 370a, 370b) receiving the same input signal as the plurality of series circuits, and means including a single pole switch (374) for selectively applying the further path signal summed with the output signal of the inverting means to the main path to boost or buck the main path signal.
17. A circuit according to claim 15 further comprising a second further path having its input coupled to the input or the output of the main path and its output coupled to the combining circuit, and including therein means for generating an equalization signal, which, at least in a part of the frequency band, boosts or bucks the main path signal.
18. A combination of circuits according to any of claims 1 through 15 wherein each circuit operates only in a portion of the overall frequency band, the combination of the circuits providing modification of the dynamic range of the input signal throughout the overall frequency band.
19. The combination of circuits according to claim 18 wherein there are two circuits (250, 252) operating, respectively, in the upper part of the frequency band and in the lower part of the frequency band.
20. The combination of circuits according to claim 19 wherein the circuits are for use at audio frequencies and the circuit (250) operating in the upper part of the frequency band operates mainly above 800 Hz



and the circuit (252) operating in the lower part of the frequency band operates mainly below 800 Hz.

21. A dual-path circuit comprising
 - a main path (254) which is linear with respect to dynamic range,
 - a combining circuit (258) in the main path, and
 - a further path including
 - at least two circuits according to any of claims 1 through 14, the inputs of the circuits being connected in parallel and constituting the input of the further path, and
 - summing means (256) for summing together the output signals from the at least two circuits, the output of the summing means constituting the output of the further path,
 - the further path having its input coupled to the input of the main path (254) or to the output of the main path and its output coupled to the combining circuit (258), the further path providing a signal which, at least in a part of the frequency band, boosts or bucks the main path signal by way of the combining circuit, but which is so limited that, in the upper part of the input dynamic range, the further path signal is limited to a value smaller than the main path signal.
22. A dual-path circuit according to claim 21 further comprising a second further path having its input coupled to the input or the output of the main path and its output coupled to the combining circuit, and including therein means for generating an equalization signal which, at least in a part of the frequency band, boosts or bucks the main path signal.
23. A dual-path circuit according to claim 21 wherein each of the at least two circuits operates only in a portion of the overall frequency band, the at least two circuits providing modification of the dynamic range of the input signal throughout the overall frequency band.
24. The dual-path circuit according to claim 23 wherein there are two circuits (250, 252) in the further path, the two circuits operating, respectively, in the upper part of the frequency band and in the lower part of the frequency band.
25. The dual-path circuit according to claim 24 for use at audio frequencies, wherein the circuit (250) operating in the upper part of the frequency band operates mainly above 800 Hz and the circuit (252) operating in the lower part of the frequency band operates mainly below 800 Hz.
26. A dual-path circuit arrangement having an overall main path and an overall further path wherein the overall further path includes a series arrangement of at least one dual-path circuit (340-342) according to claims 15 or 21, and a circuit according to any of claims 1 through 14 (344), said circuit (344) being the last in the series, the dual-path circuit arrangement including overall combining means (356) for summing the outputs of the further path(s) (250, 252) of the at least one dual-path circuit (340, 342) and the output of the last circuit (344) to provide the overall further path signal of the dual path circuit.
27. A dual-path circuit arrangement according to claim 26 further comprising means including a single pole switch for selectively applying the overall further path signal to boost or buck the overall main path signal.
28. A dual-path circuit arrangement according to claim 27 further comprising a second further path (350, 352) having its input coupled to the input or the output of the overall main path and its output coupled to the overall combining circuit (356), and including therein means (352) for generating an equalization signal, which, at least in a part of the frequency band, boosts or bucks the overall main path signal.
29. The combination of circuits according to claim 19, comprising a plurality of dynamic action stages (284-294) connected in series, the stages having staggered thresholds, at least two of said stages including first and second substages (250, 252),
 - the first and second substages each having a single pole band determining filter (154, 234),
 - the band determining filter (154) in each of said first substages having substantially the same high pass frequency characteristic and corner frequency, and
 - the band determining filter (234) in each of said second substages having substantially the same low pass frequency characteristic and corner frequency,
 - all of said corner frequencies being substantially the same,



whereby the overall effect of the series stages provides a quasi-multi-pole effect.

30. The combination of circuits according to claim 29 wherein each of said stages (284-294) comprises a dual path circuit, said dual-path circuit including a main path which is linear with respect to dynamic range, a combining circuit (318-328) in the main path, two further paths comprising said first (250) and second (252) substages having their inputs coupled to the input or output of the main path and their outputs coupled to the combining circuit (318-328) the further paths providing a signal which, at least in a part of the frequency band, boosts or bucks the main path signal by way of the combining circuit, but which is so limited that, in the upper part of the input dynamic range, the further path signal is limited to a value smaller than the main path signal for most signal conditions.

Patentansprüche

1. Schaltung zur Modifizierung des Dynamikbereichs eines Eingangssignals in einem Frequenzband, umfassend
- eine Vielzahl von Schaltungselementen (14-20; 30-36; 38-44; 58-62, 66), von denen jedes Schaltungselement eine durch $t_1(s)$, $t_2(s)$, ..., $t_n(s)$ repräsentierte Spannungsübertragungskennlinie aufweist, wobei n die Anzahl der Schaltungselemente ist, die Übertragungskennlinie frequenzabhängig oder den Dynamikbereich modifizierend oder sowohl frequenzabhängig als auch den Dynamikbereich modifizierend ist,
- wenigstens eines der Schaltungselemente wenigstens eine den Dynamikbereich modifizierende Übertragungskennlinie aufweist und
- die Schaltungselemente Übertragungskennlinien aufweisen, die wenigstens teilweise innerhalb desselben Frequenz- und Pegelbereichs arbeiten,
- eine Einrichtung (22-28; 46-56; 64) zur Verbindung der Schaltungselemente untereinander derart, daß, wenn ein durch die Spannung V_{in} dargestelltes Eingangssignal angelegt wird, die Ausgangsspannung V_{out} dargestellt wird durch

$$V_{out} = V_{in} \left[t_1(s) + \sum_{x=2}^n t_x(s) \left(\prod_{y=1}^{x-1} (1 - t_y(s)) \right) \right]$$

2. Schaltung nach Anspruch 1, bei der die Einrichtung zur Verbindung der Schaltungselemente untereinander die Schaltungselemente (14-20) derart untereinander verbindet, daß das Ausgangssignal der Schaltung tatsächlich die Summe der einzelnen Ausgangssignale der Schaltungselemente (22) ist und das tatsächliche Eingangssignal zu jedem Schaltungselement ein Differenzsignal ist, das von dem Eingangssignal und den einzelnen Ausgangssignalen von einem oder mehreren der anderen Schaltungselemente oder einem Bezugspotential abgeleitet ist.
3. Schaltung nach Anspruch 1, bei dem die Schaltungselemente (30-36) jeweils von einem Netzwerk mit drei Anschlüssen dargestellt werden, von denen jeder erste Anschluß und jeder zweite Anschluß Eingangsanschlüsse und jeder dritte Anschluß ein Ausgangsanschluß ist, die Übertragungskennlinie von dem zweiten Anschluß zum dritten Anschluß das Komplement der Übertragungskennlinie vom ersten Anschluß zum dritten Anschluß ist, das Eingangssignal auf den ersten Anschluß jedes der Schaltungselemente gekoppelt wird, die Einrichtung zur Verbindung der Schaltungselemente untereinander die Schaltungselemente in einer Anordnung untereinander verbindet, bei der der zweite Anschluß eines ersten Schaltungselements (30) mit einem Bezugspotential verbunden ist und der dritte Anschluß jedes Schaltungselements (30-34) mit dem zweiten Anschluß eines folgenden Schaltungselements (32-36) verbunden ist und das Ausgangssignal von dem dritten Anschluß des letzten Schaltungselements (36) abgenommen wird.
4. Schaltung nach Anspruch 3, bei der die Einrichtung zur Verbindung der Schaltungselemente untereinander, die Schaltungselemente in einer Anordnung untereinander verbindet, die wenigstens einen Zweig enthält derart, daß der dritte Anschluß eines Schaltungselements (58) mit dem zweiten Anschluß



von zwei oder mehr Schaltungselementen (60, 62) verbunden ist, bei denen ihrerseits das Ausgangssignal von ihrem dritten Anschluß abgenommen wird, oder ihr dritter Anschluß mit dem zweiten Anschluß von einem oder mehr Schaltungselementen (66) verbunden ist.

- 5 5. Schaltung nach einem der Ansprüche 1 - 4 bei der die Schaltung zum Komprimieren des Dynamikbereichs des Eingangssignals ist,
 wenigstens zwei Schaltungselemente (126, 128) vorhanden sind, die eine den Dynamikbereich modifizierende Übertragungskennlinie aufweisen, wobei es sich bei der den Dynamikbereich modifizierenden Übertragungskennlinie um eine den Dynamikbereich reduzierende Kennlinie handelt, und
 10 die Schaltung eine charakteristische Wirkungssubstitution liefert derart, daß die Übertragung von Komponenten des Eingangssignals mit sehr niedrigem Pegel durch die Schaltung dazu neigt, maximiert zu werden.
6. Schaltung nach einem der Ansprüche 1 - 4, bei der
 15 die Schaltung zum Expandieren des Dynamikbereichs des Eingangssignals ist,
 wenigstens zwei Schaltungselemente (126, 128) vorhanden sind, die eine den Dynamikbereich modifizierende Übertragungskennlinie aufweisen, wobei die den Dynamikbereich modifizierende Übertragungskennlinie eine den Dynamikbereich vergrößernde Kennlinie ist, und
 20 die Schaltung eine charakteristische Wirkungssubstitution liefert derart, daß die Übertragung von Komponenten des Eingangssignals mit sehr niedrigem Pegel durch die Schaltung dazu neigt, minimiert zu werden.
7. Schaltung nach einem der Ansprüche 1 - 6, bei der wenigstens zwei Schaltungselemente (126, 128) mit einer den Dynamikbereich modifizierenden Übertragungskennlinie vorhanden sind, wobei die den Dynamikbereich modifizierende Übertragungskennlinie auf den Pegel und/oder den Spektralgehalt von Signalkomponenten in dem jeweiligen Schaltungselement oberhalb eines Schwellenwerts anspricht.
8. Schaltung nach Anspruch 7, bei der zwei Schaltungselemente (126, 128) vorhanden sind, von denen jedes eine den Dynamikbereich modifizierende Übertragungskennlinie aufweist, eines der Schaltungselemente (126) ein Festbandschaltungselement ist, das den Dynamikbereich innerhalb eines festen Frequenzbandes modifiziert, und das andere der Schaltungselemente (128) ein Gleitbandschaltungselement ist, das den Dynamikbereich innerhalb eines gleitenden Frequenzbandes modifiziert.
9. Schaltung nach Anspruch 8, bei der das Gleitbandschaltungselement (129) eine gyrator-simulierte Komponente (139) enthält, die naturgemäß unerwünschtes Rauschen erzeugt, wobei die Schaltung ferner eine Einrichtung (157) umfaßt, um die gyrator-simulierte Komponente während Signalzuständen unwirksam zu machen, bei denen die gyrator-simulierten Komponente die Dynamikwirkung der Schaltung nicht beeinflußt.
- 40 10. Schaltung nach Anspruch 8 oder 9, bei der das Festbandschaltungselement (126) eine Steuerschaltung (136) enthält und die Schaltung ferner eine Einrichtung (153, 155) aufweist, um die Steuerschaltung auf einem Pegel gerade unterhalb ihres Schwellenwertes zu halten, wenn der Pegel und der Spektralgehalt des Eingangssignals derart sind, daß sich das Festbandschaltungselement unterhalb seines Schwellenwerts befindet.
- 45 11. Schaltung nach Anspruch 8, 9 oder 10, bei der die den Dynamikbereich modifizierende Übertragungskennlinie des Gleitbandschaltungselements einen Ruhezustand aufweist und die Schaltung ferner eine Einrichtung (151, 153) aufweist, um die Kennlinie von ihrem Ruhezustand wegzubewegen, wenn der Pegel und der Spektralgehalt des Eingangssignals derart sind, daß sich das Festbandschaltungselement unterhalb seines Schwellenwertes befindet.
- 50 12. Schaltung nach einem der Ansprüche 7 - 11, ferner umfassend eine Einrichtung (150) zum Koppeln von Signalkomponenten zwischen oder unter den Schaltungselementen, die eine den Dynamikbereich modifizierende Übertragungskennlinie aufweisen, damit der Schwellenwert einer oder mehrerer der Schaltungselemente beeinflußt wird.
- 55 13. Schaltung nach Anspruch 12, bei der die Einrichtung zur Kopplung von Signalkomponenten eine Filtereinrichtung (152) zur Begrenzung des Frequenzbandes der gekoppelten Signalkomponenten



einschließt.

14. Schaltung nach den Ansprüchen 12 oder 13, bei der die Schaltungselemente, die eine den Dynamikbereich modifizierende Übertragungskennlinie aufweisen, eine Einrichtung zur Ableitung eines Steuersignals (136, 146) einschließen, wobei die Einrichtung zur Kopplung von Signalkomponenten (150) eine Einrichtung zur differentiellen Ableitung des Steuersignals innerhalb eines Schaltungselements derart einschließt, daß das Steuersignal von dem Signalausgang eines anderen Schaltungselements oder den Signalausgängen anderer Schaltungselemente im wesentlichen unbeeinflusst bleibt.
15. Schaltung nach einem der Ansprüche 1 - 14, bei der es sich um eine Zweiwegeschaltung handelt, bei der die Vielzahl von Schaltungselementen und die Einrichtung zur Verbindung der Schaltungselemente untereinander einen weiteren Weg (250) bilden, die Schaltung ferner einen Hauptweg (254) aufweist, der in bezug auf den Dynamikbereich linear ist und der Hauptweg eine Kombinierschaltung (258) enthält, wobei der Eingang des weiteren Wegs mit dem Eingang des Hauptwegs (254) oder dem Ausgang des Hauptwegs (263) verbunden ist, während sein Ausgang mit der Kombinierschaltung (258) verbunden ist, der weitere Weg ein Signal liefert, das wenigstens in einem Teil des Frequenzbandes das Hauptwegsignal mittels der Kombinierschaltung unterstützt oder ihm entgegen wirkt, das aber so begrenzt ist, daß im oberen Teil des Eingangsdynamikbereichs das Signal des weiteren Wegs auf einen Wert kleiner als der des Hauptwegsignals begrenzt ist.
16. Schaltung nach Anspruch 15, bei dem der weitere Weg eine Vielzahl von Reihenschaltungen (284-288) enthält, wobei von den Reihenschaltungen wenigstens eine eine Zweiwegeschaltung ist, die Schaltung ferner eine Invertiereinrichtung (370, 370a, 370b) enthält, die dasselbe Eingangssignal wie die Vielzahl von Reihenschaltungen empfängt, und eine einen einpoligen Schalter (374) enthaltende Einrichtung zum wahlweisen Anlegen des mit dem Ausgangssignal der Invertiereinrichtung summierten Signals des weiteren Wegs an den Hauptweg aufweist, um das Hauptwegsignal zu unterstützen oder ihm entgegenzuwirken.
17. Schaltung nach Anspruch 15, ferner umfassend einen zweiten weiteren Weg, dessen Eingang mit dem Eingang oder dem Ausgang des Hauptwegs verbunden ist und dessen Ausgang mit der Kombinierschaltung verbunden ist, und in der eine Einrichtung zur Erzeugung eines Entzerrungssignals enthalten ist, daß wenigstens in einem Teil des Frequenzbandes das Hauptwegsignal unterstützt oder ihm entgegenwirkt.
18. Kombination von Schaltungen nach einem der Ansprüche 1 - 15, bei der jede Schaltung lediglich in einem Teil des Gesamtfrequenzbandes arbeitet, wobei die Kombination der Schaltungen eine Modifizierung des Dynamikbereichs des Eingangssignals im Gesamtfrequenzband liefert.
19. Kombination von Schaltungen nach Anspruch 18, bei der zwei Schaltungen (250, 252) vorhanden sind, die in dem oberen Teil des Frequenzbandes, bzw. dem unteren Teil des Frequenzbandes arbeiten.
20. Kombination von Schaltungen nach Anspruch 19, bei der die Schaltungen zur Verwendung bei Audiofrequenzen ausgelegt sind und die Schaltung (250), die im oberen Teil des Frequenzbandes arbeitet, hauptsächlich oberhalb von 800 Hz arbeitet, während die Schaltung (252), die im unteren Teil des Frequenzbandes arbeitet, hauptsächlich unterhalb von 800 Hz arbeitet.
21. Zweiwegeschaltung, umfassend
 - einen Hauptweg (254) der in bezug auf den Dynamikbereich linear ist,
 - eine Kombinierschaltung (258) in dem Hauptweg, und
 - einen weiteren Weg, enthaltend
 - wenigstens zwei Schaltungen, nach einem der Ansprüche 1 - 14, wobei die Eingänge dieser Schaltungen parallelgeschaltet sind und den Eingang des weiteren Weges bilden, und
 - eine Summiereinrichtung (256) zum Aufsummieren der Ausgangssignale der wenigstens zwei Schaltungen, wobei der Ausgang der Summiereinrichtung den Ausgang des weiteren Wegs darstellt,
 - wobei der Eingang des weiteren Wegs mit dem Eingang des Hauptwegs (254) oder dem Ausgang des Hauptwegs verbunden ist, während sein Ausgang mit der Kombinierschaltung (258) verbunden ist, wobei der weitere Weg ein Signal liefert, das wenigstens in einem Teil des Frequenzbandes, mittels der Kombinierschaltung das Hauptwegsignal unterstützt oder ihm entgegenwirkt, aber so begrenzt ist,



daß im oberen Teil des Eingangsdynamikbereichs, das Signal des weiteren Wegs auf einen Wert begrenzt ist, der kleiner als der des Hauptwegsignals ist.

22. Zweiwegeschaltung nach Anspruch 21, ferner umfassend einen zweiten weiteren Weg, dessen Eingang mit dem Eingang oder dem Ausgang des Hauptwegs verbunden ist und dessen Ausgang mit der Kombinierschaltung verbunden ist, und in dem eine Einrichtung zur Erzeugung eines Entzerrungssignals enthalten ist, das wenigstens in einem Teil des Frequenzbandes das Hauptwegsignal unterstützt oder ihm entgegenwirkt.
23. Zweiwegeschaltung nach Anspruch 21, bei der jede der wenigstens zwei Schaltungen nur in einem Teil des Gesamtfrequenzbandes arbeitet, wobei die wenigstens zwei Schaltungen eine Modifizierung des Dynamikbereichs des Eingangssignals über den gesamten Frequenzband liefern.
24. Zweiwegeschaltung nach Anspruch 23, bei der in dem weiteren Weg zwei Schaltungen (250, 252) vorhanden sind und die zwei Schaltungen in dem oberen Teil des Frequenzbandes bzw. dem unteren Teil des Frequenzbandes arbeiten.
25. Zweiwegeschaltung nach Anspruch 24 zur Verwendung bei Audiofrequenzen, bei der die Schaltung (250), die in dem oberen Teil des Frequenzbandes arbeitet, hauptsächlich oberhalb von 800 Hz arbeitet, während die Schaltung (252), die in dem unteren Teil des Frequenzbandes arbeitet, hauptsächlich unterhalb von 800 Hz arbeitet.
26. Zweiwegeschaltungsanordnung mit einem Gesamthauptweg und einem Gesamt-Weiteren-Weg, bei der der Gesamt-Weiterer-Weg eine Reihenanordnung von wenigstens einer Zweiwegeschaltung (340 - 342) nach den Ansprüchen 15 oder 21 und einer Schaltung gemäß einem der Ansprüche 1 - 14 (344) enthält, wobei die Schaltung (344) die letzte in der Reihe ist und die Zweiwegeschaltungsanordnung eine Gesamtkombiniereinrichtung (356) zum Summieren der Ausgangssignale des weiteren Wegs (der weiteren Wege) (250, 252) der wenigstens einen Zweiwegeschaltung (340, 342) und des Ausgangssignals der letzten Schaltung (344) zur Lieferung des Signals des Gesamt-Weiteren-Wegs der Zweiwegeschaltung, enthält.
27. Zweiwegeschaltungsanordnung nach Anspruch 26, ferner umfassend eine einen einpoligen Schalter einschließende Einrichtung zum wahlweisen Anlegen des Signals des Gesamt-Weiteren-Wegs, um das Signal des Gesamthauptwegs zu unterstützen oder ihm entgegenzuwirken.
28. Zweiwegeschaltungsanordnung nach Anspruch 27, ferner umfassend einen zweiten weiteren Weg (350, 352), dessen Eingang mit dem Eingang oder dem Ausgang des Gesamthauptwegs verbunden ist und dessen Ausgang mit der Gesamtkombinierschaltung (356) verbunden ist, und in dem eine Einrichtung (352) zur Erzeugung eines Entzerrungssignals enthalten ist, das wenigstens in einem Teil des Frequenzbandes das Hauptwegsignal unterstützt oder ihm entgegenwirkt.
29. Kombination von Schaltungen nach Anspruch 19, umfassend eine Vielzahl von in Reihe geschalteten Dynamikwirkungsstufen (284-294), die gestaffelte Schwellenwerte aufweisen, wobei wenigstens zwei der Stufen eine erste und eine zweite Unterstufe (250, 252) enthalten, wobei die erste und die zweite Unterstufe jeweils ein einpoliges, bandbestimmendes Filter (154, 234) aufweisen,
das bandbestimmende Filter (154) in jeder der ersten Unterstufen im wesentlichen dieselbe Hochpaßfrequenzcharakteristik und Eckfrequenz aufweisen und
das bandbestimmende Filter (234) in jeder der zweiten Unterstufen im wesentlichen dieselbe Tiefpaßfrequenzcharakteristik und Eckfrequenz aufweisen und
alle die Eckfrequenzen im wesentlichen gleich sind, und
wobei die Gesamtwirkung der Reihenstufen einen quasi Vielpoleffekt liefert.
30. Kombination von Schaltungen nach Anspruch 29, bei der jede der Stufen (284-294) eine Zweiwegeschaltung umfaßt, die Zweiwegeschaltung einen Hauptweg enthält, der in bezug auf den Dynamikbereich linear ist, eine Kombinierschaltung (318-328) in dem Hauptweg, zwei weitere Wege, die die erste (250) und die zweite (252) Unterstufe aufweisen und deren Eingänge mit dem Eingang oder Ausgang des Hauptwegs und deren Ausgänge mit der Kombinierschaltung (318-328) verbunden sind, wobei die



weiteren Wege ein Signal liefern, das wenigstens in einem Teil des Frequenzbandes mittels der Kombinierschaltung das Hauptwegsignal unterstützt oder ihm entgegenwirkt, aber so begrenzt ist, daß im oberen Teil des Eingangsdynamikbereichs das Signal des weiteren Wegs auf einen Wert begrenzt ist, der kleiner als das Hauptwegsignal für die meisten Signalzustände ist.

5

Revendications

1. Circuit pour la modification de la plage dynamique d'un signal d'entrée dans une bande de fréquences, comprenant :

- 10 - une pluralité d'éléments de circuit (14 à 20; 30 à 36; 38 à 44; 58 à 62, 66);
 - chaque élément de circuit possédant une caractéristique de transfert de la tension représentée par $t_1(s)$, $t_2(s)$, ..., $t_n(s)$, où n est le nombre desdits éléments de circuit, ladite caractéristique de transfert dépendant de la fréquence ou modifiant la plage dynamique ou dépendant, à la fois, de la fréquence et modifiant la plage dynamique;
 15 - au moins un desdits éléments de circuit possédant au moins une caractéristique de transfert modifiant la plage dynamique;
 - lesdits éléments de circuit possédant des caractéristiques de transfert fonctionnant au moins partiellement dans la même zone de fréquences et de niveaux;
 20 - des moyens (22 à 28; 46 à 56; 64) pour raccorder les éléments de circuit de telle façon que, lorsqu'un signal d'entrée représenté par la tension V_{in} est appliqué, la tension de sortie V_{out} est représentée par :

25

$$V_{out} = V_{in} \left[t_1(s) + \sum_{x=2}^n t_x(s) \left(\prod_{y=1}^{x-1} (1 - t_y(s)) \right) \right]$$

- 30 2. Circuit selon la revendication 1, dans lequel ledit moyen de raccordement des éléments de circuit raccorde lesdits éléments de circuit (14 à 20) de telle façon que la sortie du circuit soit effectivement la somme des sorties individuelles desdits éléments de circuit (22) et l'entrée effective de chaque éléments de circuit est un signal différentiel dérivé dudit signal d'entrée et les sorties individuelles d'un ou de plusieurs des autres éléments de circuit ou un potentiel de référence.

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3. Circuit selon la revendication 1, dans lequel lesdits éléments de circuit (30 à 36) sont représentés chacun comme un circuit à trois bornes dans lequel chaque première borne et chaque seconde borne sont des bornes d'entrée et chaque troisième borne est une borne de sortie, la caractéristique de transfert de la seconde borne vers la troisième borne étant complémentaire de la caractéristique de transfert de la première borne à la troisième borne, le signal d'entrée étant couplé à la première borne de chacun desdits éléments de circuit, ledit moyen de raccordement des éléments de circuit raccordant les éléments de circuit en un agencement dans lequel la seconde borne d'un premier élément de circuit (30) est couplée à un potentiel de référence et la troisième borne de chaque élément de circuit (30 à 34) est couplée à la seconde borne d'un éléments de circuit successif (32 à 36) et la sortie est dérivée de la troisième borne du dernier éléments de circuit (36).

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4. Circuit selon la revendication 3, dans lequel ledit moyen de raccordement des éléments de circuit raccorde les éléments de circuit en un agencement comprenant au moins un branchement de telle façon que la troisième borne d'un élément de circuit (58) soit couplée à la seconde borne de deux ou de plusieurs éléments de circuit (60, 62) qui, à leur tour, possèdent une sortie dérivée de leur troisième borne ou leur troisième borne couplée à la seconde borne d'un ou de plusieurs éléments de circuit (66).

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5. Circuit selon l'une quelconque des revendications 1 à 4, dans lequel :
- 55 - le circuit est prévu pour comprimer la gamme dynamique du signal d'entrée;
 - il y a au moins deux éléments de circuit (126, 128) possédant une caractéristique de transfert modifiant la plage dynamique, la caractéristique de transfert modifiant la plage dynamique étant une caractéristique de réduction de la plage dynamique; et



- le circuit assure une substitution d'action caractéristique telle que la transmission de composantes de très bas niveau du signal d'entrée à travers le circuit ait tendance à être maximisée.
6. Circuit selon l'une quelconque des revendications 1 à 4, dans lequel :
- le circuit est prévu pour l'expansion de la gamme dynamique du signal d'entrée;
 - il y a au moins deux éléments de circuit (123, 128) possédant une caractéristique de transfert modifiant la plage dynamique, la caractéristique de transfert modifiant la plage dynamique étant une caractéristique de diminution de la plage dynamique; et
 - le circuit fournit une substitution d'action caractéristique tel que la transmission de composantes de très bas niveau du signal d'entrée à travers le circuit ait tendance à être minimisée.
7. Circuit selon l'une quelconque des revendications 1 à 6, dans lequel on trouve au moins deux éléments de circuit (126, 128) possédant une caractéristique de transfert modifiant la plage dynamique, la caractéristique de transfert modifiant la plage dynamique étant sensible au contenu de niveau et/ou spectral des composantes du signal dans l'élément de circuit respectif au dessus d'un seuil.
8. Circuit selon la revendication 7, dans lequel on trouve deux éléments de circuit (126, 128) possédant chacun une caractéristique de transfert modifiant la plage dynamique, un desdits éléments de circuit (126) étant un élément de circuit de bande fixe modifiant la plage dynamique dans une bande de fréquences fixée et l'autre desdits éléments de circuit (128) étant un élément de circuit de bande mobile modifiant la plage dynamique dans une bande de fréquences mobile.
9. Circuit selon la revendication 8, dans lequel l'élément de circuit de bande mobile (129) comprend un élément bobiné et simulé (139) qui génère, de façon inhérente, un bruit indésirable, le circuit comprenant, de plus, un moyen (157) pour rendre le composant bobiné simulé inefficace dans des conditions du signal lorsque le composant bobiné simulé n'affecte pas l'action dynamique du circuit.
10. Circuit selon les revendications 8 ou 9, dans lequel l'élément de circuit de bande fixe (126) comprend un circuit de commande (136) et le circuit comprend de plus des moyens (1253, 155) pour maintenir ledit circuit de commande à un niveau juste en dessous de son niveau de seuil lorsque le contenu de niveau et spectral du signal d'entrée est tel que l'élément de circuit de bande fixe est en dessous de son seuil.
11. Circuit selon les revendications 8, 9 ou 10, dans lequel la caractéristique de transfert modifiant la plage dynamique de l'élément de circuit de bande mobile possède une condition au repos, le circuit comprenant, de plus, des moyens (151, 153) pour déplacer ladite caractéristique à l'écart de sa condition au repos lorsque le contenu de niveau spectral du signal d'entrée est tel que l'élément de circuit de bande fixe est en dessous de son seuil.
12. Circuit selon l'une quelconque des revendications 7 à 11, comprenant, de plus, un moyen (150) pour le couplage des composantes du signal entre ou parmi lesdits éléments de circuit possédant une caractéristique de transfert modifiant la plage dynamique pour affecter le niveau de seuil d'un ou de plusieurs desdits éléments de circuit.
13. Circuit selon la revendication 12, dans lequel ledit élément de couplage des composantes du signal comprend un moyen de filtre (152) pour restreindre la bande de fréquences des composantes du signal couplées.
14. Circuit selon les revendications 12 ou 13, dans lequel lesdits éléments de circuit possédant une caractéristique de transfert modifiant la plage dynamique comprennent des moyens de dérivation d'un signal de commande (136, 146), ledit moyen de couplage des composantes du signal (150) comprenant un moyen pour dériver, de façon différentielle, le signal de commande dans un élément de circuit de façon à ce que le signal de commande soit pratiquement non affecté par la(les) sortie(s) de signal d'un autre élément de circuit ou d'autres éléments de circuit.
15. Circuit selon l'une quelconque des revendications 1 à 14, dans lequel le circuit est un circuit à double accès dans lequel ladite pluralité d'éléments de circuit et ledit moyen de raccordement des éléments de circuit constituent un accès supplémentaire (250), le circuit comprenant, de plus, un accès principal



- (254) linéaire par rapport à la plage dynamique, un circuit de combinaison (258) dans l'accès principal, ledit accès supplémentaire voyant son entrée couplée à l'entrée de l'accès principal (254) ou à la sortie de l'accès principal (263) et sa sortie couplée au circuit de combinaison (258), l'accès supplémentaire fournissant un signal qui, au moins dans une partie de la bande de fréquence, amplifie ou s'oppose au signal d'accès principal à l'aide du circuit de combinaison mais qui est limité de façon à ce que, dans la partie supérieure de la plage dynamique d'entrée, le signal d'accès supplémentaire soit limité à une valeur plus petite que le signal d'accès principal.
16. Circuit selon la revendication 15, dans lequel l'accès supplémentaire comprend une pluralité de circuits en série (284 à 288), au moins un desdits circuits en série étant un circuit à double accès, le circuit comprenant de plus des moyens d'inversion (370, 370a, 370b) recevant le même signal d'entrée que la pluralité des circuits en série et des moyens comprenant un commutateur à pôle unique (374) pour l'application sélective du signal d'accès supplémentaire sommé avec le signal de sortie du moyen d'inversion à l'accès principal pour amplifier ou pour s'opposer au signal d'accès principal.
17. Circuit selon la revendication 15, comprenant, de plus, un second accès supplémentaire dont l'entrée est couplée à l'entrée ou à la sortie de l'accès principal et dont la sortie est couplée au circuit de combinaison et comprenant un moyen pour la génération d'un signal d'égalisation qui, au moins dans une partie de la bande de fréquences, amplifie ou s'oppose au signal d'accès principal.
18. Combinaison de circuits selon l'une quelconque des revendications 1 à 15, dans laquelle chaque circuit ne fonctionne que dans une partie de la bande de fréquences globale, la combinaison des circuits assurant une modification de la plage dynamique du signal d'entrée sur la bande de fréquences globales.
19. Combinaison de circuits selon la revendication 18, dans laquelle on trouve deux circuits (250, 252) fonctionnant respectivement dans la partie supérieure de la bande de fréquence et dans la partie inférieure de la bande de fréquences.
20. Combinaison de circuits selon la revendication 19, dans laquelle les circuits sont prévus pour des fréquences audio et le circuit (250) fonctionnant dans la partie supérieure de la bande de fréquences fonctionne principalement au dessus de 800 Hz et le circuit (252), fonctionnant dans la partie inférieure de la bande de fréquences, fonctionne principalement en dessous de 800 Hz.
21. Circuit à double accès comprenant :
- un accès principal (254) linéaire par rapport à la plage dynamique;
 - un circuit de combinaison (258) dans l'accès principal et un accès supplémentaire comprenant :
 - au moins deux circuits selon l'une quelconque des revendications 1 à 14, les entrées des circuits étant montées en parallèle et constituant l'entrée de l'accès supplémentaire; et
 - un moyen de sommation (256) des signaux de sortie à partir d'au moins deux circuits, la sortie du moyen de sommation constituant la sortie de l'accès supplémentaire;
- l'accès supplémentaire voyant son entrée couplée à l'entrée de l'accès principal (254) ou à la sortie de l'accès principal et sa sortie couplée au circuit de combinaison (258), l'accès supplémentaire fournissant un signal qui, au moins dans une partie de la bande de fréquences, amplifie ou s'oppose au signal d'accès principal au moyen du circuit de combinaison mais qui est limité de façon à ce que, dans la partie supérieure de la plage dynamique d'entrée, le signal d'accès supplémentaire soit limité à une valeur plus faible que le signal d'accès principal.
22. Circuit à double accès selon la revendication 21, comprenant, de plus, un second accès supplémentaire dont l'entrée est couplée à l'entrée ou à la sortie de l'accès principal et dont la sortie est couplée au circuit de combinaison et comprenant un moyen pour la génération d'un signal d'égalisation qui, au moins dans une partie de la bande de fréquences, amplifie ou s'oppose au signal d'accès principal.
23. Circuit à double accès selon la revendication 21, dans lequel chacun d'au moins deux circuits ne fonctionne que dans une partie de la bande de fréquences globales, les au moins deux circuits assurant une modification de la plage dynamique du signal d'entrée sur la bande de fréquences globale.



24. Circuit à double accès selon la revendication 23, dans lequel on trouve deux circuits (250, 22) dans l'accès supplémentaire, les deux circuits fonctionnant respectivement dans la partie supérieure de la bande de fréquences et dans la partie inférieure de la bande de fréquences.
- 5 25. Circuit à double accès selon la revendication 24, pour des fréquences audio, dans lequel le circuit (250) fonctionnant dans la partie supérieure de la bande de fréquences fonctionne principalement au dessus de 800 Hz et le circuit (252) fonctionnant dans la partie inférieure de la bande de fréquences fonctionne principalement en dessous de 800 Hz.
- 10 26. Agencement de circuit à double accès possédant un accès principal global et un accès supplémentaire global, dans lequel l'accès supplémentaire global comprend un agencement en série d'au moins un circuit à double accès (340 à 342) selon les revendications 15 ou 21 et un circuit (344) selon l'une quelconque des revendications 1 à 14, ledit circuit (344) étant le dernier de la série, l'agencement de circuit à double accès comprenant un moyen de combinaison global (356) pour la sommation des
15 sorties du (des) accès supplémentaire(s) (250, 252) d'au moins le circuit à double accès (340, 342) et la sortie du dernier circuit (344) pour fournir le signal global d'accès supplémentaire du circuit à double accès.
- 20 27. Agencement de circuit à double accès selon la revendication 26, comprenant, de plus, un moyen comprenant un commutateur à pôle unique pour l'application sélective du signal global d'accès supplémentaire afin d'amplifier ou de s'opposer au signal global d'accès principal.
- 25 28. Agencement de Circuit à double accès selon la revendication 27, comprenant de plus un second accès supplémentaire (350, 352) dont l'entrée est couplée à l'entrée ou à la sortie de l'accès principal global et dont la sortie est couplée au circuit de combinaison global (356) et comprenant un moyen (352) pour la génération d'un signal d'égalisation qui, au moins dans une partie de la bande de fréquences, amplifie ou s'oppose au signal global d'accès principal.
- 30 29. Combinaison de circuits selon la revendication 19, comprenant une pluralité d'étages à action dynamique (284 à 294) montés en série, les étages possédant des seuils étagés, au moins deux desdits étages comprenant des premier et second sous-étages (250, 252), les premier et second sous-étages possédant chacun un filtre de détermination de bande à pôle unique (154, 234), le filtre de détermination de bande (154) dans chacun desdits premiers sous-étages présentant pratiquement la même caractéristique de fréquence passe-haut et la même fréquence de basculement, et le filtre de
35 détermination de bande (234) dans chacun desdits seconds sous-étages possédant pratiquement la même caractéristique de fréquence passe-bas et la même fréquence de basculement, toutes lesdites fréquences de basculement étant pratiquement la même, l'effet global des étages en série assurant ainsi un effet de quasiment plusieurs pôles.
- 40 30. Combinaison de circuits selon la revendication 29, dans laquelle chacun desdits étages (284 à 294) comprend un circuit à double accès, ledit circuit à double accès comprenant un accès principal linéaire par rapport à la plage dynamique, un circuit de combinaison (318 à 328) dans l'accès principal, deux accès supplémentaires comprenant lesdits premier (250) et second (252) sous-étages dont les entrées sont couplées à l'entrée ou à la sortie de l'accès principal et dont les sorties sont couplées au circuit
45 de combinaison (318 à 328), les accès supplémentaires fournissant un signal qui, au moins dans une partie de la bande de fréquence, amplifie ou s'oppose au signal d'accès principal au moyen de circuit de combinaison mais qui est limité de façon à ce que, dans la partie supérieure de la plage dynamique d'entrée, le signal d'accès supplémentaire soit limité à une valeur plus faible que le signal d'accès principal pour la plupart des conditions du signal.
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