

Description

1. Field of the Invention

The present invention is in the field of amplifiers, and relates particularly to amplifiers which drive loads such as speakers that are highly reactive and are also subject to mechanical distortion influences including inertia and resonances.

2. Description of the Prior Art

For more than forty years, and still according to the current state of the art, audio amplifiers have employed what is commonly referred to as "voltage feedback" in an endeavor to improve frequency response and reduce distortion. Such voltage feedback systems are sometimes referred to in the art as "constant voltage" systems, since for a fixed amplifier input voltage the output voltage remains substantially constant over a broad frequency range or bandwidth. Thus, present audio amplifiers are capable of providing a voltage output for driving a speaker which quite accurately follows the amplifier input program voltage, as to both waveform or shape and phase.

However, the conventional magnetic coil-driven speaker (or multiple speaker system) used as a load for such amplifier is essentially a current-driven device, and it has both electrical characteristics and mechanical characteristics which seriously alter the flow of current through it independently of "constant voltage" amplifier drive, thereby preventing the speaker current from coming even close to following the voltage output of the amplifier and instead causing the speaker to depart considerably from the program applied by the amplifier in amplitude, waveform and phase. The general result is that the acoustic response of the speaker is considerably different than the flat voltage response of the "constant voltage" amplifier.

The conventional coil-driven speaker (or multiple speaker system) is a load that is highly inductively reactive. This causes the load impedance to vary with frequency, becoming higher, with power to the speaker consequently lower, as program frequencies rise above the usual 400 Hz nominal or rated impedance point all of the way out to 20 KHz and higher. By way of example, to illustrate how serious this increase in impedance at higher frequencies can be, the manufacturer's impedance/frequency response curve for a typical state-of-the-art speaker rated 8 ohms at 400 Hz shows the impedance to be doubled to 16 ohms at approximately 4300 Hz, and quadrupled to 32 ohms at about 10KHz. Power is inversely proportional to impedance in a speaker.

The inductive reactance of the speaker load

also causes load current to lag in phase from the program, and this phase lag increases with frequency from the 400 Hz rated impedance point all the way to the high end of the sound frequency spectrum. By way of example to show how serious this inductive reactance phase lag can be, measurements made with the speaker referred to in the immediately preceding paragraph showed a phase lag of approximately 33° at 400 Hz, approximately 45° at 900 Hz, and approximately 70° at 5 KHz. As an integral part of this inductive reactance phase lag, the rise times for high frequency wave fronts or transients which usually contain important program information are slowed way down relative to the actual wave fronts or transients in the program.

The mass of a speaker resists acceleration and deceleration in response to respective rising and falling wave fronts, resulting in inertial lag and overshoot, respectively. Inertial distortion of speaker acoustic output is surprisingly close or analogous to the phase and rise time distortions produced by the inductive reactance of a speaker, so that these effects of inductive reactance and inertia are additive. The adverse effects of inertia, like those of inductive reactance, increase with increasing frequency all of the way out to the high end of the audible spectrum.

Most sounds produced by musical instruments have a sharp attack characterized by a sharply rising initial transient wave front in each fundamental frequency cycle, this initial transient containing most of the high frequency harmonic content of the sound. It has been found that for the human ear to hear the entire spectrum of such sounds, it must receive these initial high frequency harmonic sounds first, followed then by the midrange and low end frequencies. However, the additive or cumulative effects of speaker inductive reactance and inertia in current state-of-the-art amplifier/speaker system cause the rise time to be so slow and the phase lag to be so large at higher frequencies that the sharply rising wave fronts or initial transients which contain most of the high frequency harmonics become masked to a large extent by the lower, heavier frequencies. Such masking of the high frequency harmonics is commonly referred to as "transient distortion," and causes the acoustic output of the system to sound "artificial" or "recorded," instead of sounding completely "live" or "natural" as when the ear properly receives the sharply rising initial transient wave front in its proper order ahead of the lower frequencies.

The typical condenser microphone has a rise time of about 20 to 25 microseconds. In order to make up for the losses at the sound reproduction end, and at the same time reproduce the high frequency harmonic content in the sharply rising initial transient wave fronts captured by such a

condenser microphone, it is necessary for an amplifier system to be able to produce a speaker load current rise time that is faster than the microphone rise time, preferably about 10 microseconds or less, and ideally about 5 microseconds or less. Direct solid state pickup transducers, such as those manufactured and sold by Barcus-Berry, Inc. of Huntington Beach, California, introduce virtually no delay into the program rise time and where such direct transducers are employed, if the amplifier system has a rise time of not more than about 10 microseconds, and preferably not more than about 5 microseconds, the reproduction will sound completely "live" and natural." In order to produce such rapid rise times in an amplifier/speaker combination, the speaker load current phase must be held close to "in phase" with the program signal from the 400 Hz nominal impedance point all of the way out to approximately 20 KHz, and preferably the speaker load current should be leading from approximately 1 KHz all of the way out to approximately 20 KHz. Additionally, in order for the high frequency harmonics to be heard in their proper proportion relative to the other sound components, the attenuation of these high frequency components resulting from speaker inductive reactance must also be overcome.

Applicant has discovered that by deliberately providing the high frequency harmonics and transients with a large phase lead of as much as 180° relative to the midrange frequencies, the resulting separation of the highs from the midrange frequencies enables the highs to be heard by the ear before the midrange, resulting in the highs being heard in much better detail.

Conventional state-of-the-art amplifier/speaker systems are not capable of providing speaker load current with the fast rise time and substantially "in phase" condition required to avoid serious masking problems. Various attempts have been made to solve the problem, and while some of these have improved the rise time and others have improved the phasing, none have heretofore simultaneously produced a rise time on the order of 10 microseconds or less and an "in phase" or leading phase condition of the high frequency harmonics and transients all of the way out to about 20 KHz. Additionally, applicant is not aware of any previous attempts to unmask the high frequencies from the midrange frequencies by providing the high frequencies with a large phase lead of as much as 180° from the midrange frequencies so as to improve the high frequency detail heard. A principal type of equipment currently employed in attempting to solve this problem is the graphic equalizer. Thus, a quality unit in the hi-fi industry is the Bose equalizer marketed with the Bose speaker. This unit has a frequency response curve that is actually

too steep and a slightly leading phase up to around 10 KHz, going in phase between about 10 KHz and 15 KHz, but then the response curve falls off rapidly, and there is a large phase lag after 15 KHz, the lag being approximately 45° at 20 KHz. The resulting rise time is about 45 to 40 microseconds, much too slow to reproduce the high frequency harmonics in the initial transient wave front of most musical instrument sounds.

Below the 400 Hz nominal impedance point, the impedance curve for a conventional speaker rises in a capacitive reactance effect caused by the compliance and open air cone resonance of the speaker. The manufacturer's speaker response curve referred to above shows the impedance to rise sharply below the 150 Hz up to a large impedance peak at 50 Hz, and then to slope back down sharply to 8 ohms at 20 Hz. This large low frequency impedance peak represents a big hole in the acoustic output of the speaker, so that much of the low frequency information in the program is lost. While some cabinet designs are effective to reduce open air cone resonance at low frequencies, they generally introduce further problems such as speaker cabinet resonances and undesired damping. Unless amplifier compensation can be provided to raise up the low frequency response all of the way down to about 20 Hz, much of the low frequency content will be lost, and particularly the low frequency information representing the percussion sounds.

In addition to the poor low frequency response of the typical speaker, the capacitive reactance effect in the region of the rising slope of the cone resonance part of the response curve below 400 Hz produces a seriously leading phase, causing fundamental and low harmonic frequencies in this region, in effect, to overtake the simultaneously phase-lagging high frequency components of the program, further compounding the "masking" problem referred to above.

The most pertinent prior art of which applicant is aware is US-A-4,260,954 "Amplifier Load Correction System." The load correction system of this prior patent developed feedback voltage signal representative of current through the driven load, made a comparison between such feedback signal and the amplifier input program voltage, and utilized the results of this comparison to adjust the gain of the program amplifier line to compensate for load current deviations in waveform and phase from the program. This prior system worked well when it sensed the load current of a single speaker, but attempts to employ it in connection with a load that was more complex than a single speaker, such as a plurality of speakers with crossover networks or a multiple speaker distribution line system, were not always satisfactory because the

more complex load currents did not necessarily truly represent speaker performance, and the sensing could thus be masked. Another problem with direct sensing of the real load as in US-A-4,260,954 was that the real load was unpredictable and often imperfect, having undesirable impedance humps and dips relative to frequency. The prior system was not capable of straightening out such irregularities. The system of US-A-4,260,954 also had the difficulty that it could not be simply plugged into the amplification line; it also required additional connections to the driven load.

US-A-4,260,954 disclosed no means for automatically adjusting the average amplitude of the high frequencies relative to the average amplitude of the midrange frequencies to compensate for program imbalance between these frequency ranges. Also, US-A-4,260,954 did not disclose the concept of deliberately introducing a leading phase to the high frequencies so as to separate them and unmask them from the midrange frequencies.

DE-C-2,235,664, was one of the references cited in US-A-4,260,954, the Figure 2 circuit of that reference being relied upon. However, such circuit as shown and described in this German patent does not suggest use of a reference or model load to correct for the adverse effects of inductive reactance, inertia and resonances of the real speaker load. In fact, the circuit of the German patent does not in any way tend to correct for deviations of speaker load current or acoustic output from the incoming program.

Other references relied upon in US-A-4,260,954 which may be of interest relative to the present invention because of a rough general similarity in circuit appearance to circuits in one or more of the forms of the present invention are US-A-3,902,111 and US-A-4,153,849. However, these two references are considered to constitute nonanalogous art, in that they are not dealing with any kind of a source of a program voltage signal that is variable as to waveform, or with the typical amplifier load such as speaker that is intended to be driven in response to such a variable waveform program voltage signal and which conventionally has a load current and power output that varies widely and generally continuously from the program voltage signal in both waveform and phase due to its high reactance, inertia, and resonances. Thus, the circuit shown and described in US-A-3,902,111 appears to be simply a low frequency servo system, with a fixed set point input and no way to put in a program voltage signal variable as to waveform, and with type and values of circuit components completely unrelated to and leading away from the present invention.

US-A-4,153,849 relates to a circuit for normalizing the operating characteristics of yttrium-iron-

garnet (YIG) devices and traveling-wave maser (TWM) devices tuned by a controlled current so that such devices, despite variations in individual ones, can be interchanged in a microwave system without realigning the entire system. Again, no program signal variable as to waveform is applicable to the input, but there is simply a reference DC voltage which is applied to shift the response of the device to the required current-frequency characteristic for the prealigned overall system. US-A-4,153,849 discloses a positive feedback circuit arrangement having type and values of circuit components completely inapplicable to and leading away from the present invention.

US-A-4,052,560 discloses a method of correcting for variations in the load with the features set forth in the first part of claim 1, and a load correction system with the features included in the first part of claim 10. The known system includes an integrator which produces a very gradual response slope from the low end toward the high end, and a differentiator which produces a very gradual response from the high end toward the low end, and these two response curves are quite flat within the radio range from 20 Hz to 20 kHz, theoretically averaging out to a generally flat frequency response. Neither the integrator nor the differentiator is capable of producing any appreciable phase shift. Therefore, neither of these circuit elements will provide a material correction for the inductive phase lag of a speaker voice coil, which increases throughout the audio frequency range.

SUMMARY OF THE INVENTION

In view of the above problems in the art, it is a general object of the present invention to provide a system which substantially completely corrects for adverse characteristics of an amplifier-driven load that would otherwise cause the output of the load to deviate from the applied program, such adverse characteristics including electrical resistance, inertia, resonances, and the like.

A specific object of the invention resides in the provision of a method and system of correcting for variations in a load, in which HF components are delivered with a phase lead with respect to midrange frequencies, thereby enabling a better separation of the HF components from the midrange frequencies and unmasking the HF components.

This object is met by the invention as characterized in claims 1 and 10.

The invention has particular utility in audio systems, substantially completely correcting for the normally seriously distorting effects of inductive reactance, inertial lag and overshoot, and speaker compliance and associated open air cone reso-

nance.

The system described below provides an amplifier load correction system capable of so completely overcoming the usual phase lagging characteristics of a speaker which increase from the nominal or rated 400 Hz frequency all of the way out to 20 KHz caused by both inductive reactance and inertia, that for the first time in the art the speaker load current phase can be held substantially completely "in phase," and even slightly leading in phase, from 400 Hz all of the way out to 20 KHz, while simultaneously the rise time can be kept below about 10 microseconds in all forms of the invention, and within about a 2 to 5 microsecond range for some forms of the invention. Another related object of the invention is to provide an amplifier load correction system capable of providing a load current rise time that is sufficiently rapid and has sufficient amplitude, together with the simultaneous maintenance of a substantially completely "in phase" condition from approximately 400 Hz all of the way out to approximately 20 KHz, to enable the amplifier-speaker system to faithfully reproduce the sharply rising initial transient fronts containing most of the high frequency harmonics in musical instrument and some voice frequency sounds, and delivering this sharply rising initial transient wave front and then the midrange and low end frequencies to the ear in the correct sequential order, thereby completely overcoming transient distortion masking, and enabling the reproduced sounds for the first time to be perceived as completely "live" and "natural" in form, rather than sounding "artificial" or "recorded" as from prior art systems.

In the amplifier load correction system of the character described, a large leading phase of as much as 180° is deliberately applied to the high frequencies relative to the midrange frequencies so as to set the high frequency harmonics and transients out to the ear before the onset of hearing the midrange frequencies, and thereby substantially completely unmask the high frequencies from the midrange frequencies.

The amplifier load correction system is capable of providing a low frequency lagging load current phase so as to compensate for the normally leading phase of a speaker in the region of the rising slope of the cone resonance part of the speaker response curve below 400 Hz, such low frequency phase compensation further assisting in the avoidance of masking of the high frequency harmonics by low frequency sound components.

The amplifier load correction system is further capable of producing a load current rise time considerably faster than the 20 to 25 microsecond rise time of the typical condenser microphone, so as to make up for losses at the sound production end,

and at the same time reproduce the high frequency harmonic content in the sharply rising initial transient wave fronts captured by such a condenser microphone.

The amplifier load correction system, while providing fast rise time and phase corrections, at the same time corrects for the normally poor speaker impedance curve, and hence power output response, both in the high frequency range from 400 Hz all the way out to 20 KHz and in the low frequency range from 400 Hz all the way down to below 20 Hz.

The amplifier load correction system develops the correction signal from a reference load that serves as a model of the characteristics of the real load, such reference load being completely isolated and independent of the real driven load. The reference load may serve as an equivalent model for the inductance of the real speaker load, and also serve as an analog model for the mechanical inertia of the speaker load, the system providing a sufficient correction amplitude to simultaneously correct for the usual adverse effects of both the inductive reactance and the inertia of the real speaker or speaker system. Similarly, the reference load may also serve as an equivalent model for the capacitance of the real speaker system load and at the same time as an analog model for the compliance and related open air cone resonance of the speaker system, to correct for these speaker characteristics.

By deriving its correction signal from a reference or model load instead of deriving the correction signal from the real driven load, the system enables the correction to be based upon a substantially perfect or ideal load, so that the corrected transducer output can be a substantially perfect reproduction of the program. The correction signal is not limited or distorted by unpredictable imperfections such as impedance humps and dips in the real load.

The correction system may be employed in connection with multiple speaker systems, such as a plurality of speakers with crossover networks or a multiple speaker distribution line system, whereas such multiple speaker systems might provide a confused basis for correction in prior systems which compare current through the real driven load with the program signal.

The amplifier load correction system is particularly easy to hook up in any amplifier-speaker system, requiring only an input jack from the preamplifier and an output jack to the power amplifier, and not requiring the additional connections to the driven load as were required in systems which sensed load current in the real driven load.

In the reference load amplifier correction system of the character described, the reference load

includes inductor and capacitor components, or circuits simulating such components, which are scaled up to have much larger values than the corresponding inductance and capacitance characteristics of the real driven load represented by the reference load, to minimize current through the reference load and hence power required to drive the reference load.

In the amplifier load correction system of the character described, the high frequencies, midrange frequencies and low frequencies are separated out into three separate channels so that the amplitudes can be adjusted relative to each other for generally flat performance from approximately 20 Hz all of the way up to 20 KHz. An automatic balancing system between the high frequency and midrange frequency channels is provided so that, if the average amplitude of the high frequencies increases or decreases relative to the average amplitude of the midrange frequencies, the gain of the high frequency channel will automatically be respectively reduced or increased to bring the amplitudes of the high and midrange frequency channels into desired balance despite imbalances that may occur in the program.

In an embodiment of the invention, program signal is applied to a reference load having both inductor and capacitor components, or circuits simulating such components, so as to develop respective high and low frequency correction signal components. These inductor and capacitor components (or simulations thereof) of the reference load are preferably tuned to cancel and provide zero correction signal from the system at approximately the 400 Hz nominal impedance frequency of most speakers and speaker systems, the inductor component (or simulation thereof) of the reference load developing a high frequency correction signal for frequencies from approximately 400 Hz all of the way out to approximately 20 KHz, and the capacitor component (or simulation thereof) of the reference load developing a low frequency correction signal for frequencies from approximately 400 Hz down to below about 20 Hz.

Since circuits simulating an inductor or a capacitor can be employed in place of actual inductors and capacitors as reference load components in the various forms of the present invention, for convenience and conciseness in the following description of the invention and in the appended claims, the terms "inductor component," "inductor component means," "capacitor component" and "capacitor component means" are hereby defined to mean both real and simulated inductor and capacitor components.

In some forms of the invention, the reference load is a single series-tuned inductor-capacitor circuit from which the high and low frequency correc-

tion signal components are developed together. In other forms of the invention, separate high and low frequency channels are employed, the high frequency channel containing the inductor reference load component and the low frequency channel containing the capacitor reference load component. The correction signal outputs of the high and low frequency channels are mixed with the program signal in a summing amplifier for delivery to the power amplifier that drives the real load. This use of separate high and low frequency correction signal channels, with a further independent program signal channel, enables the amplitudes of the high and low frequencies to be independently adjusted relative to the program signal and each other, so that the signals may be delivered in optimum relative proportions to the power amplifier driving the real load.

In some forms of the invention, the inductor and capacitor components in respective high and low frequency channels are simplified RC (resistor and capacitor) phase shifting networks which shift the phases of the high and low frequency portions of the audio spectrum in the appropriate directions and are at the same time amplitude responsive to frequency, so as to provide substantially complete compensation for adverse characteristics of the load resulting from inductive and capacitive reactances and load inertia. This is particularly useful in the high frequency channel because it enables a small, inexpensive RC circuit-type inductor component to be employed in lieu of a real choke, thereby permitting the undesirable bulk and high cost of a real choke inductor to be eliminated from the system, and enabling the load correction circuit of the invention to be embodied in a small hybrid IC (integrated circuit) chip.

In one form of the invention in which there are separate high, midrange and low frequency channels, amplitude imbalances are detected between the average amplitude of the high frequencies and average amplitude of the midrange frequencies, and the average amplitude of the high frequencies is automatically readjusted to re-establish a balance between the high and midrange frequencies. Such automatic readjustment is accomplished by a detector and summing system wherein a negative DC output detector senses the signal in the high frequency channel and produces a DC voltage analog thereof, a positive DC output midrange detector senses the signal in the midrange channel and produces a DC analog thereof, and the minus and plus DC outputs of these detectors are mixed in a summing amplifier, the DC output of which automatically adjusts a variable resistor to readjust the gain of the high frequency channel as required.

In one form of the invention which has separate high, midrange and low frequency channels, one of

the high and midrange channels has an inverter therein which gives the high frequencies a 180° phase lead over the midrange frequencies, setting the high frequency harmonics and transients out to the ear before the onset of hearing the midrange frequencies and thereby substantially completely unmasking the high frequencies from the midrange frequencies.

In another form of the invention, adjustment of the correction signal relative to the program signal is provided for by having two channels, the correction signal containing both high and low frequency correction components being developed in one channel in which the program signal is cancelled so that the correction signal can be independently adjusted, and the other channel being a direct program signal channel. The adjusted correction signal and the program signal are mixed in a summing amplifier for delivery to the power amplifier that drives the real load.

In some forms of the invention, current through the reference load is sensed and a voltage feedback signal representative of reference load current is compared with the program signal voltage in a differential operational amplifier (or in separate differential operational amplifiers if separate high and low frequency correction channels are employed) to develop the correction signal that is to be delivered to the power amplifier along with the program signal. In other forms of the invention, the inductor component of the reference load is arranged, and preferably each of the inductor and capacitor components of the reference load is arranged in a separate channel voltage divider network from which the correction signal component is derived; and the correction signals, which are independently adjustable, are mixed with the program signal in a summing amplifier for delivery to the power amplifier that drives the real load.

BRIEF DESCRIPTION OF THE DRAWINGS

These and other objects of the invention will become more apparent in view of the following description taken in conjunction with the drawings, wherein:

Figure 1 is a circuit diagram of a load correction system in which separate high and low frequency channels contain respective separate inductor and capacitor reference load components.

Figure 2 is a circuit diagram illustrating another form of the system also having separate channels, but in which the high frequency channel correction signal is developed in a divider network containing an active inductor.

Figure 3 is a circuit diagram illustrating another form of the system which has separate high and

low frequency channels in which the respective high and low frequency correction signals are derived from voltage dividers containing the respective inductor and capacitor reference load components.

Figure 4 is a circuit diagram illustrating a further form of the system which has separate high, midrange and low frequency channels and in which the inductor and capacitor reference load components in the respective high and low frequency channels are simplified RC phase shifting networks.

Figure 5 is a circuit diagram illustrating an embodiment of the invention which also has separate high, midrange and low frequency channels, but which additionally has a circuit system which automatically readjusts the amplitude of the high frequency signal relative to that of the midrange signal to maintain a desired amplitude balance, and which has a phase inverter in the midrange channel that gives the high frequency channel signal a 180° phase lead over the midrange channel signal.

Figure 6 is a block diagram, with some circuit components illustrated, of the circuit diagram shown in Figure 5, but with the addition of a recording stage between the reference load correction system of the present invention and the power amplifier, for recording the program and correction signals for later simultaneous applications to the power amplifier.

Figure 7 is a correction signal amplitude vs. frequency curve for the correction signal of the present invention from 20 Hz to 20 KHz.

Figure 8 shows a total impedance vs. frequency response curve for a typical speaker that is rated 8 ohms at 400 Hz.

DETAILED DESCRIPTION

Referring to the drawings, Figure 1 is a circuit diagram illustrating a first form of a reference load amplifier correction circuit or system of the present invention, shown in Figure 6 in operative association with a power amplifier and speaker load which may be any existing, conventional state-of-the-art power amplifier and speaker components. Although such power amplifier and associated load are not shown in Figures 1-5, it is to be understood that the power amplifier and load will be connected to the output terminals of each of these circuit forms.

The reference load correction circuit or system shown in Figure 6 is generally designated 10, and includes input and output terminals 12 and 14, respectively. Input terminal 12 is electrically connected to a program source, which will generally include a conventional state-of-the-art preamplifier.

The correction circuit output terminal 14 is connected to the input terminal 16 of power amplifier 18 which may be a conventional state-of-the-art power gain block. To assure adequate amplifier power is available for the reference load correction circuit 10 of the invention to substantially completely conform the output of the load transducer to the waveform and phase of the program input to correction circuit 10, over a wide frequency band, it is preferred that the power amplifier 18 have a power gain of at least approximately 20. Additionally, in order to assure that the power amplifier does not introduce any material waveform distortion, it is preferred that its frequency response be substantially flat (in the constant voltage mode) from approximately 20 Hz to approximately 20 KHz.

While the reference load correction circuit 10 of the invention will normally be connected between a preamplifier and a power amplifier, it is to be understood that it may be disposed at any location in the line of components from the program source to the power amplifier.

While the reference load correction circuit of the present invention is capable of correcting for the defects or deficiencies of any amplifier-driven load, the typical load will be a speaker load such as that found in a hi-fi, television or radio set. The speaker load may be a single speaker, such as the speaker 20 diagrammatically illustrated in Figure 6, or may consist of multiple speakers with crossover networks. The reference load correction circuit of the invention is also suitable for load correction in a multiple-channel amplifier system where the channels are bridged together for a monaural output of multiplied power. The reference load correction circuit of the invention is also available for load correction in a distribution line system (e.g., 70 volts) to a number of speakers, as sometimes employed in theaters. In all multiple speaker systems where the reference load correction circuit of the invention is employed, the circuit of the invention will substantially completely correct for both the electrical (reactance) and mechanical (inertial) deficiencies of each of the plurality of speakers in the system.

The program signal furnished to the correction circuit input terminal 12 is a voltage signal variable as to waveform according to the music, voice or other program information which it contains. Under conventional circumstances, without the presence of the reference load correction circuit or system 10 of the invention, after amplification in the constant voltage power amplifier 18, this variable waveform program signal would be subjected to severe amplitude and phase distortions by the high inductive reactance, the mechanical inertia, and the open air cone resonance of the speaker device load 20, as discussed in detail hereinafter in connection with the load correction characteristics of

the invention.

Referring now to Figure 1, the reference load correction system shown in this figure includes separate high frequency and low frequency reference load correction circuit portions 118 and 120, respectively, each of which has its own respective balanced differential operational amplifier 122 and 124. The program input terminal 12d is connected, through input resistor 126, to the noninverting input of each of the differential amplifiers 122 and 124. The high frequency differential amplifier 122 uses an inductor component reference load 128 and the low frequency differential amplifier 124 uses a capacitor component reference load 130, with the crossover frequency for the reference loads 128 and 130 still at 400 Hz. The output 132 of amplifier 122 is connected to one side of the inductor component reference load 128, while the other side of reference load 128 is connected to sensing point 134, which is in turn connected both to sensing resistor 136 and to the inverting input of amplifier 122. Similarly, the output 138 of amplifier 124 is connected to one side of the capacitor component reference load 130, the other side of which is connected to sensing point 140, the sensing point 140 in turn being connected to sensing resistor 142 and to the inverting input of amplifier 124.

Separate high and low frequency correction signals are provided at the respective amplifier outputs 132 and 138, and these are mixed through respective individual gain control potentiometers 144 and 146 with the original program into a summing amplifier 104a. The outputs of the respective high and low frequency correction signal gain control potentiometers 144 and 146 are connected through respective summing resistors 148 and 150 to a summing bus 152 that is connected to the inverting input 106a of summing amplifier 104a. Flat, full, undistorted program signal is conducted from program input terminal 12d through a conductor 110a and summing resistor 154 to the summing bus 152. Thus, the separately adjustable high and low frequency correction signals are mixed with the original program signal in the summing bus 152 and summing amplifier 104a. Preferably, the feedback resistor 114a connected between the output 116a and inverting input of summing amplifier 104a has the same value as each of the summing resistors 148, 150 and 154 for a one-to-one gain for each of the mixed signals. The output 116a of summing amplifier 104a is connected to the system output 14d, which in turn is adapted to be connected to a power amplifier and load combination.

The circuit of Figure 1, while containing the desirable feature of independent adjustability of the high and low frequency corrections, has a rapid rise time and slightly leading phase all the way out

to 20 KHz. The harmonic distortion is reduced to approximately .4 percent.

The circuit illustrated in Figure 2 has, as does the form of Figure 1, three separate channels, high and low frequency correction channels, and a straight-through program signal channel. In the form of Figure 2, the low frequency correction channel and the program signal channel are the same as in the circuit of Figure 1, but the high frequency channel employs as the inductor component an "active inductor" in a voltage divider circuit arrangement that produces the high frequency amplitude and phase correction signal without requiring the preparation of a feedback signal or the comparison of such a feedback signal with the incoming program signal as required for the high frequency correction in the previously described forms of the invention.

In Figure 2, the program voltage signal is introduced at program input terminal 12e and is fed to separate high and low frequency reference load correction circuit portions 118a and 120a, respectively. As in Figure 1, the low frequency correction circuit portion 120a includes a balanced differential operational amplifier 124a which drives a capacitor component reference load 130a to develop the required low frequency voltage feedback signal across sensing resistor 142a and deliver this feedback voltage signal to the inverting input of the differential amplifier 124a. The low frequency correction signal is delivered from the output of differential amplifier 124a through gain control potentiometer 146a and summing resistor 150a to summing bus 152a.

As in the circuit of Figure 1, the flat, unattenuated program voltage signal is delivered from program input terminal 12e through conductor 110b and summing resistor 154a to the summing bus 152a.

The high frequency reference load correction circuit portion 118a in Figure 2 utilizes as the inductor component an active inductor in a voltage divider circuit arrangement which produces a high frequency correction signal essentially the same as the high frequency correction signal produced in Figure 1, but is produced without the Figure 1 feedback and comparison with the incoming program signal. The active inductor is shown enclosed in the dotted line block and is generally designated 156. A conventional inductor could be employed in the same circuit position as the active inductor 156 in Figure 2, but the active inductor is preferred because it functions as a perfect inductor that has no resistance. The active inductor 156 is a type of circuit well known in the art, so the detailed operation of its individual parts does not form a part of the present invention. It includes back-to-back differential operational amplifiers 158 and 160, the

amplifier 160 having phase shifting networks 162 and 164 associated with its respective inverting and noninverting inputs, the phase shifting network 162 consisting of a resistor 166 and capacitor 168, and the phase shifting network 164 consisting of a capacitor 170 and resistor 172. The active inductor amplifier 160 has an output resistor 174.

The voltage divider of which the active inductor 156 is a part includes a resistor 176 as the upper part of the voltage divider, the resistor 176 being connected at its upper end to program input terminal 12e and at its lower end to the active inductor 156 at a junction 178. The active inductor 156 is the lower part of the voltage divider, extending from the junction 178 to a ground connection 180. The output from this voltage divider consisting of resistor 176 and active inductor 156 is taken from a junction 182 at the output of amplifier 158. The output point 182 will correspond in amplitude and phase to the voltage divider junction point 178, because the amplifier 158 is arranged as a unity gain, noninverting voltage follower.

The amplitude of the high frequency correction signal is made adjustable by being taken off of the junction point 182 through gain control potentiometer 144a, and the high frequency correction signal is delivered through a summing resistor 148a to the summing bus 152a. The summing bus 152a delivers the high and low frequency correction signals and the program signal to summing amplifier 104b, the output 116b of which is connected to the system output terminal 14e that is adapted to be connected to a power amplifier and load combination.

The gain on the three channels in the summing amplifier 104b is the value of the amplifier feedback resistor 114b divided by the value of the respective channel summing resistor 148a for the high frequency channel, 150a for the low frequency channel, and 154a for the program channel. Inasmuch as the voltage divider arrangement of the high frequency channel causes its correction signal to be considerably lower in amplitude than the correction signal of the low frequency and program channels, the summing resistor 148a for the high frequency channel has a much lower value than the summing resistors for the low frequency and program signal channels, as for example in a ratio of one to ten, so as to provide a balanced output between the three channels.

The active inductor high frequency channel reference load operates in the following manner to provide the high frequency correction signal. At the 400 Hz crossover frequency, the active inductor 156 of the high frequency channel 118a and the capacitor component reference load 130a of the low frequency channel 120a balance so as to cancel. As program frequency rises above the 400 Hz

crossover frequency, the active inductor 156 increases more and more in impedance, so that it occupies an increasingly larger proportion of the voltage drop in the voltage divider arrangement with resistor 176, to produce an increasingly higher correction signal voltage output at the junction point 182 and hence through the gain control potentiometer 144a and summing resistor 148a to the summing bus 152a and summing amplifier 104b. The continuous, instantaneous nature of the correction signal, with its fast rise time that is faster than the audible harmonics, causes the high frequency correction signal output from the voltage divider output junction point 182 to have a slightly leading phase relative to the program signal all of the way up from the 400 Hz crossover frequency to 20 KHz. The phase of the high frequency correction signal is a leading phase, being essentially the continuous, instantaneous complement of the phase lag caused by the active inductor 156.

The reference load correction circuit of Figure 2 has exhibited a much lower harmonic distortion than the previously described correction circuits, on the order of .1 percent, improved rise time of from about 2 to about 3 microseconds, and does not cause any change in phase correction or rise time with adjustments of the amount of correction in the respective high and low frequency channels 118a and 120a.

In the reference load correction system illustrated in Figure 3, separate high and low frequency channels 118b and 120b are employed as they are in the systems of Figures 1 and 2, but the reference loads for the separate high and low frequency channels are each arranged in a voltage divider configuration similar to the high frequency channel voltage divider configuration of Figure 2. If desired, the high frequency correction channel 118b of the Figure 3 system can be made identical to the high frequency correction channel 118a of Figure 2, employing the active inductor 156 of Figure 2. However, to illustrate the high frequency channel inductor component voltage divider arrangement in a simpler form, a real inductor 156a has been shown as the inductor component in Figure 3 instead of the active inductor 156.

The high frequency channel voltage divider in Figure 3 consists of resistor 176a and inductor component 156a connected between program input terminal 12f and ground. The high frequency correction signal output is from the junction 178a between the voltage divider components 176a and 156a, and is fed through gain control potentiometer 144b and summing resistor 148b to summing bus 152b and thence to summing amplifier 104c. This high frequency correction channel 118b of Figure 3 operates in exactly the same way as the high frequency correction channel 118a of Figure 2.

The low frequency correction channel 120b of Figure 3 includes a voltage divider between program input terminal 12f and ground consisting of a resistor 184 and a reference load capacitor component 186. The output junction 188 from this reference load divider provides its low frequency correction signal through gain control potentiometer 146b and summing resistor 150b to summing bus 152b and thence to summing amplifier 104c.

The flat, unattenuated program signal is delivered from program input terminal 12f through conductor 110c and summing resistor 154b to summing bus 152b and thence to summing amplifier 104c. The mixed high and low frequency correction signals and program signal are delivered from the output 116c of summing amplifier 104c to the system output 14f which is adapted to be connected to a power amplifier and load combination.

Because of attenuation in the voltage dividers of both the high frequency correction channel 118b and the low frequency correction channel 120b, the respective summing resistors 148b and 150b for the high and low frequency channels are made much smaller than the feedback resistor 114c in summing amplifier 104c and the summing resistor 154b in the program channel conductor 110c. This causes the correction signals from both the high frequency correction channel 118b and the low frequency correction channel 120b to be amplified to correct proportions relative to the program signal.

The reference load inductor component 156a in high frequency correction channel 118b and the reference load capacitor component 186 in the low frequency correction channel 120b are balanced to cancel at the 400 Hz crossover frequency. The capacitor component reference load voltage divider in the low frequency channel 120b operates in the same manner as the inductor component reference load voltage dividers in the high frequency channels of both Figure 2 and Figure 3, except that it operates in the opposite frequency direction for frequency reductions below 400 Hz. Thus, as the frequency drops below 400 Hz, the impedance of reference load capacitor component 186 increases, so that the proportion of voltage drop across capacitor 186 increases relative to the overall voltage drop across the divider that includes resistor 184 and capacitor 186, resulting in increased correction signal output delivered from the voltage divider junction 188 to gain control potentiometer 146b and thence through summing resistor 150b and summing bus 152b to summing amplifier 104c. The phase of the low frequency correction signal produced at the voltage divider junction 188 is a lagging phase, being essentially the continuous, instantaneous complement of the phase lead caused by the capacitor component 186 as such ca-

pacitor component phase varies over the low frequency range. Thus, the leading capacitor component phase relative to the program signal will cause the low frequency correction signal to lag in phase relative to the program signal. This phase-lagging correction voltage signal will compensate for the leading phase that otherwise would be produced at low frequencies by capacitive reactance in the real load. In the same manner, but conversely, the high frequency correction voltage from high frequency channel 148b has a leading phase to compensate for the phase lag that would otherwise be caused at high frequencies by the inductive reactance of the real load.

In the load correction system of Figure 3, the input divider resistors 176a and 184 set the respective slopes of the high and low frequency portions of the correction signal curve of Figure 10. The heights of the respective high and low frequency portions of the curve are adjustably established by the settings of the respective gain control potentiometers 144b and 146b. In the system of Figure 2, the slope of the high frequency portion of the curve is set by the input divider resistor 176, while the slope of the low frequency portion of the curve is set by the sensing resistor 142a. The heights of the high and low frequency portions of the curve are adjustably set by the respective gain control potentiometers 144a and 146a. In the system of Figure 1, the slopes of the high and low frequency portions of the curve are set by the values of the respective sensing resistors 136 and 142, while the heights of the high and low sections of the curve are adjustably set by the respective gain control potentiometers 144 and 146.

The performance of the Figure 3 system is substantially the same as the performance of the Figure 2 system, with a rise time in the range of from about 2 to about 5 microseconds, a leading phase above the 400 Hz crossover frequency all the way out to 20 KHz, and the adjustments not disturbing either phase or rise time. As with the Figure 2 form, the system of Figure 3 has a very low harmonic distortion of about .1 percent. The harmonic distortion previously noted for the circuit of Figure 1 appeared to be principally a distortion of the third harmonic; however, such distortion is substantially eliminated in the Figure 2 and Figure 3 systems.

Figure 4 illustrates another simplified form of the present circuit which has separate high and low frequency channels in a manner similar to the circuits of Figures 1-3. However, the circuit of Figure 4 has several features which are improvements over the circuits of Figures 1-3. First, the inductor and capacitor reference load components in the respective high and low frequency channels of Figure 4 are simplified RC (resistor and capacitor)

phase shifting networks which shift the phases of the high and low frequency portions of the audio spectrum in the appropriate directions and are at the same time amplitude responsive to frequency generally in the same manner as the inductor and capacitor components of other forms of circuits. The RC circuit-type inductor component of the Figure 4 circuit enables the undesirable bulk and high cost of the real choke inductors of Figures 1 and 3 or the added circuit complexity of the active inductor of Figure 2 to be eliminated, and this enables the circuit of Figure 4 to be embodied in a small hybrid IC (integrated circuit) chip.

Another difference between the circuit of Figure 4 and the circuits of Figures 1-3 is that in Figure 4 the reference load inductor component in the high frequency channel and the reference load capacitor component in the low frequency channel are respective high and low pass filters, and instead of the straight-through full program signal channels 110a, 110b and 110c of Figures 5, 6 and 7, respectively, the circuit of Figure 4 has a midrange band pass filter in this same, center channel. With the high, low and midrange frequencies thus separated in the circuit of Figure 4, the highs and lows can be amplitude adjusted for the desired balance relative to the midrange. Applicant has found that excellent performance is obtained with a midrange band pass filter having its response centered at 400 Hz and with corner frequencies at approximately 700 and 200 Hz (corner frequencies being the 3 DB drop points); with the high pass filter and phase shifter rising above the midrange center level at approximately 1 KHz at a slope adjustable up to a maximum of approximately 6 DB per octave; with the low pass filter and phase shifter rising above the center level of the midrange band pass at approximately 300 Hz at a slope adjustable up to approximately 6 DB per octave and then leveling off around approximately 200 Hz and being substantially flat all the way down to below 20 Hz; and with the reference load inductor component 200 and capacitor component 210 being balanced to cancel at the 400 Hz crossover frequency. This results in a substantially flat overall response from approximately 20 Hz all of the way up to approximately 20 KHz.

Referring now to the specific details of the circuit shown in Figure 4, the high and low frequency correction channels are generally designated 118c and 120c, respectively, and they are arranged as parallel circuits between the input junction 12g and the output summing bus 152c.

The reference load inductor component in high frequency channel 118c is generally designated 200, and in this form of the invention it is a passive high pass filter and phase shifter which produces a leading phase for the high frequency part of the

spectrum, and causes both the phase lead and the amplitude of the highs to increase with frequency from approximately 1 KHz all of the way up to at least approximately 20 KHz, so as to compensate for the inductive reactance lag, frequency response characteristics and inertia characteristics of the load. This passive high pass filter and phase shifter 200 consists of a pair of resistors 202 and 204 and a pair of capacitors 206 and 208. The output of this reference load inductor component 200 is through a manual gain control potentiometer 144c and summing resistor 148c to the summing bus 152c.

The low frequency correction channel 120c has its reference load capacitor component 210 in the form of a passive low pass filter and phase shifter which produces a phase lag in the low frequency signal and causes both this phase lag and the amplitude of the low frequency signal to increase as frequency decreases from approximately 300 Hz, so as to compensate for the capacitive reactance load and frequency response characteristics of the load. This passive low pass filter and phase shifter, which is the reference load capacitor component 210, consists of a pair of resistors 212 and 214 and a pair of capacitors 216 and 218. The output of low frequency correction channel 120c is through manual gain control potentiometer 146c and summing resistor 150c to the output summing bus 152c. The reference load inductor component 200 and reference load capacitor component 210 are balanced to substantially cancel at the 400 Hz crossover frequency.

The midrange channel, generally designated 220, also is arranged as a parallel circuit with the high frequency channel 118c and low frequency channel 120c between the input junction 12g and the output summing bus 152c. Midrange channel 220 includes conductor 222 and an operational amplifier 224 in a midrange band pass filter configuration. The output of band pass filter 224 is taken through a summing resistor 154c to the output summing bus 152c.

The input to the input junction 12g is through a voltage follower 226 having a very low output impedance for impedance matching with the three parallel circuits 118c, 120c and 220. The summing bus 152c feeds into summing amplifier 104d which has feedback resistor 114d, the output of summing amplifier 104d being connected to the system output 14g. The gain of each of the three channel signals in summing amplifier 104d is the ratio of the value of feedback resistor 114d to the value of the respective channel summing resistors 148c, 150c and 154c.

The high frequency harmonics, in general, are the harmonics of fundamentals in the midrange frequency band, and the ability to adjust the amplitude of the high frequencies relative to the

midrange frequencies with the gain control potentiometer 144c in high frequency channel 118c enables an optimum balance to be achieved between the amplitude of the high frequency and the amplitude of the midrange fundamentals of such harmonics for best unmasking of the high frequency harmonics and transients. The circuit of Figure 5 provides for automatic adjustment of this balance between the highs and the midrange frequencies as this balance deviates from a desired preset balance according to variations in the program.

The performance of the Figure 4 circuit is substantially the same as the performance of the forms of circuit illustrated in Figures 2 and 3, with a rise time in the range of from about 2 to about 5 microseconds, a leading phase above the 400 Hz crossover frequency all of the way out to 20 KHz, and the adjustments in the circuit not disturbing either phase or rise time. As with the forms of Figures 2 and 3, the circuit arrangement of Figure 4 has a very low harmonic distortion of not more than about .1 percent.

Referring now to Figure 5, the input to input junction 12h is through a voltage follower 227 for impedance matching. High frequency correction channel 118d and low frequency correction channel 120d are connected in a parallel circuit arrangement between the input junction 12h and output summing bus 152d. The low frequency correction channel 120d is the same as the low frequency correction channel 120c in Figure 8, having reference load capacitor component 210 in the form of a passive low pass filter and phase shifter, with the output from low frequency correction channel 120d being through manual gain control potentiometer 146d and summing resistor 150d to the output summing bus 152d.

The reference load inductor component 228 in the high frequency correction channel 118d is an RC network that produces a phase lead which, along with amplitude, increases with frequency from approximately 1 KHz all of the way out to 20 KHz. This RC network of the high pass filter and phase shifter reference load inductor component 228 consists of a resistor 230 and a pair of capacitors 232 and 234. The output of inductor component network 228 passes through a summing or gain set resistor pair 236 and 238 which may be considered as a split resistor divided at an intermediate junction point 240. The resistor 238 is connected to the output summing bus 152d. An automatically variable resistor 242, preferably an FET (field effect transistor), is connected between such junction point 240 and ground for varying the effective resistance of the summing gain set resistor pair 236 and 238 so as to automatically balance the amplitude of the high frequencies relative to the midrange frequencies. This is to compensate auto-

matically for programs being played that are either over-equalized or under-equalized in the high frequency signal range, either during the entire program or during some portions of the program. This automatic compensation thus automatically mixes the program in the proper manner for a correct, preset balance between the high range and the midrange. Details of the automatic compensation circuit will be described hereinafter.

As in the other circuits described above, the reference load inductor component 228 and reference load capacitor component 210 are balanced to substantially cancel at the 400 Hz crossover frequency.

The midrange channel is generally designated 244, its conductor 246 being connected to the input junction 12h and leading to midrange band pass filter 248 having substantially the same frequency response characteristics as the midrange band pass 224 of Figure 8, but in this case the band pass filter 248 is also an inverter to separate the high frequency signal from the midrange signal by 180° as discussed more in detail hereinafter. The output of midrange band pass and inverter 248 goes through summing resistor 250 to the output summing bus 152d.

The automatic system in the circuit arrangement of Figure 5 for adjusting the amplitude balance between the output of high frequency correction channel 118d and midrange channel 244 is a detector and summing system wherein a negative DC output detector senses the high frequency signal in high frequency channel 118d and produces a DC voltage analog representative of the average high frequency amplitude, a positive DC output midrange detector senses the signal in midrange channel 244 and produces a DC analog output representing the average amplitude of the midrange signal, and the minus and plus DC outputs of the respective high and midrange detectors are mixed in a summing amplifier, the DC output of which automatically adjusts the variable resistor 242. A more detailed description of how this dual detector and summing amplifier circuit system works will be provided in connection with the detailed description of the circuitry itself.

The source of unattenuated midrange signal for this circuit system is a junction point 252 in the midrange channel conductor 246 at the output of midrange band pass 248. A conductor 254 leads from junction point 252 to the midrange detector 256, which is a rectifier giving an output that is an average DC of the program midrange frequencies. The positive DC output of detector 256 goes through a gain adjust potentiometer 258 which is the main adjust for the system, and then through a summing resistor 260 to an input junction 262 for a summing amplifier 264 that has a feedback resistor

266. The ratio of the value of feedback resistor 266 to the value of summing resistor 260 determines the gain which the summing amplifier 264 applies to the midrange DC. Summing amplifier 264 is negatively biased at its input junction 262 from a potentiometer 268 through a resistor 270, the amount of this negative bias varying according to the type of FET employed for the variable resistor 242; typically this will be on the order of about 3 volts.

The source of high frequency signal is a junction point 272 in the high frequency correction channel 118d between capacitor 234 and resistor 236. By placing the junction point 272 at this location, it provides a source of unattenuated high frequency signal through a conductor 273 to the high frequency detector 274, which is a negative output rectifier providing a negative DC voltage that is the analog of the average amplitude of the high frequencies. This negative DC output goes through a preset trimmer potentiometer 276 and a summing resistor 278 to the input junction 262 of DC summing amplifier 264. The gain provided to this high frequency range negative DC voltage by summing amplifier 264 is the ratio of the value of its feedback resistor 266 to the value of high frequency summing resistor 278.

The sum output of summing amplifier 264, containing positive DC information from midrange detector 256 and negative DC information from high range detector 274, passes through a conductor 280 to the FET 242 which is in a variable resistor configuration. The FET circuit is arranged so that an increase in the output voltage of summing amplifier 264 increases the resistance of variable resistor 242, while a decrease in the DC voltage output of summing amplifier 264 decreases the resistance of the variable resistor 242. The manner in which this provides for automatic adjustment of the balance between the high and midrange frequency portions of the program is as follows.

The variable resistor FET 242 forms a voltage divider to the junction point 240 in the middle of the divided gain set resistor 236-238, so that as the FET's resistance goes up and down, it allows, respectively, more or less signal to go into the primary summing amplifier 284 from the high frequency correction channel 118d. If the average amplitude of the program midrange frequencies comes up relative to the average amplitude of the program high frequencies, then the positive DC output of midrange detector 256 rises relative to the negative DC output of high range detector 274, and the DC output of summing amplifier 264 will rise accordingly and cause an increase in the resistance of variable resistor 242, increasing the gain of the high frequency channel 118d. Conversely, if

the average amplitude of the high frequency signal rises relative to the average amplitude of the midrange signal, then the negative DC output of high frequency detector 274 increases, causing a reduction in the DC output voltage of summing amplifier 264, and a resulting decrease in the resistance of the FET variable resistor 242, which decreases the gain of the high frequency channel 118d. In this manner, the system will continuously and automatically seek a proper balance between the high and midrange portions of the program, according to the settings of main adjust potentiometer 258 and high frequency preset trimmer potentiometer 276.

It will thus be seen that the midrange portion of this automatic balancing system functions in the manner of an expander, since it increases the gain of the high frequency channel 118d to match increases in the average amplitude of the midrange frequencies. Conversely, the high frequency portion of the balancing system functions in the manner of a compressor, since it decreases the gain of the high frequency channel 118d to compensate for increases in the average amplitude of the high frequency signal. Thus, the overall balancing system functions in the manner of a compander.

The automatic amplitude balancing between the midrange and high frequency signals is important because of the fact described above in connection with Figure 4 that the high frequency harmonics, in general, are the harmonics of fundamentals in the midrange frequency band, and with the high and midrange frequencies thus closely related, the amplified program is more real if they are maintained in proper balance. Although in prototype circuitry a similar automatic balancing was provided in the low frequency channel, this did not improve the amplified program output sufficiently to warrant the extra complication and expense.

The reason for the phase inversion in the midrange channel 244 provided by the midrange band pass filter and inverter 248 is to set the high frequency harmonics and transients out to the ear before the onset of hearing the midrange frequencies and thereby substantially completely unmask the high frequencies from the midrange frequencies. This is one of the characterizing features of the present invention. Thus, in the circuit arrangement of Figure 5, the high frequencies are leading in phase much further than they were in the other forms of the invention as described above, and this has been found in practice to give better separation and better details of the high frequencies from the midrange frequencies. This phase lead of the highs relative to the midrange frequencies is accomplished by inversion of the midrange channel 244 rather than inversion of the high frequency channel 118d to eliminate the need for an extra operational

amplifier since the operational amplifier 248 serves the dual purposes of band pass filter and inverter, whereas a separate operational amplifier would have been required in the high frequency channel 118d. Nevertheless, alternatively the same phase separation between highs and midrange frequencies may be accomplished by providing an inverter in the high frequency channel 118d. Although inversion of the midrange channel 244 will probably cause some separation of the low frequencies in the low frequency channel 120d, this does not seem to be audibly detectible in practice.

The correct balancing between high frequencies and midrange frequencies and the large phase lead introduced in the high frequencies relative to the midrange frequencies improves the high frequency detail so much that there may be a psychological feeling of having lost something in the base frequency range. This can be compensated for by adjusting the potentiometer 146d in low frequency channel 120d to raise the gain of the low frequency channel.

Aside from the automatic balancing between the high pass channel 118d and midrange band pass channel 244, the frequency response characteristics for all three channels are substantially the same for the circuit of Figure 5 as for the circuit of Figure 4. Also, aside from the results of the midrange channel inversion, the performance of the circuit of Figure 5 is at least as good as that of the circuits of Figures 2-4 as to rise time and low harmonic distortion.

Figure 7 shows a correction signal amplitude vs. frequency curve which is typical for the reference load correction systems illustrated in Figures 1-5.

By way of comparison, the curve shown in Figure 8 is an impedance/frequency response curve given by a manufacturer for a typical speaker rated 8 ohms at 400 Hz, this curve illustrating some of the serious problems inherent in conventional speakers that are corrected by the reference load correction circuits of the present invention.

Referring at first to Figure 7, the correction signal curve is generally designated 190, and includes high frequency portion 192 above the 400 Hz crossover frequency and low frequency portion 194 below the 400 Hz crossover frequency. At 400 Hz there is no correction signal, and both the high frequency portion 192 and the low frequency portion 194 of curve 190 rise tangentially from the 400 Hz zero point on the curve 190.

The high frequency portion 192 of curve 190 gradually slopes upwardly from the 400 Hz point, and rises between about 4 and 6 decibels per octave starting at about 1,000 Hz up through about 10 KHz, where the curve starts to roll off, the curve going nearly flat at 20 KHz. Above 20 KHz addi-

tional correction does not contribute significantly to the sound.

The low frequency portion 194 of curve 190 gradually rises below 400 Hz to a slope of about 3 decibels per octave, up to about 40 Hz, and then rolls off and levels out all of the way down to about 5 Hz.

Referring to Figure 8, the impedance curve generally designated 196 represents total speaker impedance. Above about 400 Hz this impedance is the vector sum of resistance and inductive reactance; while below about 400 Hz this impedance is the vector sum of resistance and some capacitive reactance of the speaker coil, but primarily a capacitive reactance effect caused by the compliance and open air cone resonance of the speaker. It will be noted from the total impedance curve 196 that the manufacturer's rating for the speaker of 8 ohms only holds approximately true within a very limited frequency range of from about 150 Hz to about 600 Hz, and that both above and below this very limited range, the total impedance of the speaker rises to much higher values. Thus, because of the inductive reactance of this load, its impedance doubles to 16 ohms at approximately 2300 Hz, and quadruples to 32 ohms at about 10 KHz. Below about 150 ohms, the impedance curve 196 rises sharply to an impedance peak caused primarily by the cone resonance of the speaker, and then slopes back down sharply to 8 ohms at 20 Hz.

Since power delivered to a speaker from an amplifier, and hence acoustic output of the speaker, is inversely proportional to speaker impedance, it will be apparent from the speaker impedance curve 196 of Figure 8 that above 400 Hz the higher frequencies in the amplifier program will be greatly reduced in acoustic output relative to the lower frequencies, so that much of the overtone structure will be reduced or lost, and the formant structure of the program will be considerably adversely affected. In addition to the serious amplitude losses of the program, the inductive reactance of the speaker causes large phase lags in the load current relative to the program voltage, and such phase lags greatly increase at the higher frequencies. Thus, the phase lag angle of the load current caused by the inductive reactance of the speaker represented by the impedance curve 196 is approximately 33° at 400 Hz, approximately 45° at 900 Hz, and as high as approximately 70° at 5 KHz. Such phase lag angles of the load current relative to the program signal are directly responsible for transient distortion referred to hereinabove in which the fundamental and lower harmonic frequencies mask the high frequency harmonics. The rising part of impedance curve 196 below 400 Hz caused by speaker compliance and cone resonance has a leading phase, which normally will make the mask-

ing problem even worse.

The general similarity between the correction signal curve 190 of Figure 7 provided by the present invention and the typical total speaker impedance curve 196 of Figure 8 is to be noted. The high frequency portion 192 of the correction signal curve of Figure 7 rises at a rate sufficient to substantially completely overcome the increase of impedance with frequency of the speaker as illustrated in the rising part of the speaker impedance curve 196 in Figure 8. As described in detail hereinabove, the reference load correction signal provided by the present invention substantially completely corrects for the speaker phase lags indicated in Figure 8, and the circuits of the present invention illustrated in Figures 4-8 will actually cause a slightly leading load current phase relative to the phase of the program signal voltage from the 400 Hz crossover frequency all of the way out to 20 KHz, assuring that the acoustic output of the speaker will be essentially in phase over this entire frequency spectrum. Additionally, the circuit illustrated in Figure 5 will provide a large amount of phase lead for the high frequency harmonics relative to the midrange frequencies to highlight the high frequencies, as described in detail above in connection with these Figure 5.

The low frequency portion 194 of curve 190 shown in Figure 7 does not completely compensate for the cone resonance peak, but the slope and height of the low frequency portion 194 of the curve are adjustable so that the low frequency acoustic output of the speaker or speaker system will sound good. Bringing the leveled part of the low frequency portion 194 of the curve all the way down below 20 Hz assures that all of the percussion sounds will be clearly heard. A very important aspect of the reference load correction provided by the present invention in the low frequency portion 194 of curve 190 is that the low frequency correction signal has a phase lag, on the order of about 25° to 30°, which cancels the leading phase of the rising part of the cone resonance section of the speaker impedance curve.

While the reference load correction circuit embodying the present invention has been shown and described as being adapted to correct for speaker system loads, it is to be understood that the invention is equally adaptable to correct for the deficiencies of any amplifier-driven load, with the reference load circuit components being chosen as an equivalent model for the electrical and mechanical characteristics of the driven load, whatever that load may be.

In an early experimental prototype of the invention, an actual speaker load was employed as the reference load, and in other early prototypes, a passive reference load configuration was employed

with inductor and capacitor circuit components having values selected to correspond approximately to the inductance and capacitance found in a real speaker load. However, experimentation with such early prototypes revealed that an inductor component of the reference load of the invention not only can serve as an equivalent model for the inductance of the real speaker load, but also as an analog model for the mechanical inertia of the speaker load. The current response curve in an inductor relative to the applied program voltage is closely analogous to the mechanical response curve of a speaker relative to the applied program. If a square wave voltage pulse is applied across an inductor, the current through the inductor will lag in phase at both the leading and trailing edges of the pulse; such leading edge phase lag corresponds to mechanical inertial lag, while such trailing edge phase lag corresponds to inertial overshoot. Advantage is taken of this correspondence between inductive reactance and inertia in the circuits illustrated in Figures 1-5 by providing a large enough reference load inductor component to serve as an equivalent model for the inductance of the speaker system, and additionally to serve as an analog model for the mechanical inertia of the speaker system. Thus, the inductor circuit component of the reference load is selected so that the correction signal amplitude and power are sufficient to synergistically compensate substantially completely for both the inductive reactance and the mechanical inertia of the speaker system.

Similarly, the capacitor circuit component in the reference load of the present invention has been found to be capable of serving not only as an equivalent model for the capacitance of the real speaker system load, but also as an analog model for the compliance and related open air cone resonance of the speaker system, so that the low frequency correction signal provided by the invention serves not only to compensate for the capacitive reactance of the real load speaker system, but also synergistically compensates for the compliance and cone resonance of the speaker system.

In an amplifier load correction system such as that disclosed in US-A-4,206,954 wherein corrections were made in response to sensing of the real transducer load current, where the load became more complex than a single speaker, as for example a plurality of speakers with crossover networks, or a multiple speaker distribution line system, such system could mask the sensing and interfere with the correction. The reference load correction system employed in the present invention has been found to completely overcome this problem, and experimental prototypes of the invention have been tested with a wide variety of speaker systems of varying complexity and have been found to be

completely compatible with all of them, substantially completely correcting for speaker system reactance and inertia in all cases.

Another problem in a direct load sensing system such as that of US-A-4,260,954 that is completely cured by utilizing the reference load of the present invention is that a real load is unpredictable and often imperfect, while the reference load of the present invention is completely predictable and can serve as a substantially perfect model load. Thus, while the reference load of the present invention will correct for and substantially completely cure imperfections in the real load, when an imperfect real load is sensed as in the system of US-A-4,260,954, the imperfection in the speaker system will not be corrected.

Another advantage of the reference load system of the present invention over a direct load sensing system such as that of US-A-4,260,954 is that the reference load system is much simpler to hook up in any amplifier-speaker system, requiring only an input jack from the preamplifier and an output jack to the power amplifier, and not requiring the additional connections to the driven load as were required in the direct sensing type of system.

The reference load of the present invention is completely separated and isolated from the driven load, and as indicated in the preceding comparisons between the reference load system of the present invention and a direct load sensing system such as that of US-A-4,260,954, it will be seen that this complete separation and isolation of the reference load from the real load is a critical factor in overcoming each of the problems of the direct load sensing type of system.

In a test to illustrate the phase and transient correction characteristics of the present invention, separate, identical state-of-the-art speakers were connected to each of the two channels of a single power amplifier, and a graphic equalizer was employed to zero out the frequency responses of the speakers so that the acoustic output was flat in each of the two channels. An experimental prototype of the present invention was connected between the preamplifier and power amplifier in only one of the two channels. Upon switching back and forth between the channel without the reference correction system of the invention and the channel with the reference load correction system of the invention, the audible difference was dramatic. The channel with the reference load correction system sounded just like live sound, with no masking at all, even though the program was actually a recorded program; while the channel without the reference load correction system had a "recorded" sound with much of the high frequency transient content masked.

By way of example only and not of limitation, a

type of operational amplifier that has proved to be satisfactory in the present invention is a 4558 dual operational amplifier such as that currently available as a standard chip from Motorola, National and Raytheon. This is a high performance operational amplifier with low noise and fast response characteristics. Also the TLO 62 operational amplifier from Texas Instruments has been found satisfactory. Also by way of example only and not of limitation, a part that has been found satisfactory for the variable resistor FET 242 of Figure 5 is a Siliconics J 230FET.

The various circuit component values shown in the circuit diagrams of Figures 1-5 have been found to produce satisfactory results in experimental prototypes, but these also are given by way of example only and not of limitation.

Although the present invention has been shown and described herein primarily in an arrangement in which it provides reference load correction signal directly into the amplification line which drives a load such as a speaker system, the invention may also be employed in the processing of recorded program information. Thus, the original program, whether it be live or prerecorded, is applied to an amplification line that embodies the reference load correction system of the invention, and the resulting output program that includes the correction voltage introduced by the reference load system of the invention is recorded (or re-recorded if the program came from a prior recording). Then, when the processed recording with the reference load correction signal thereon is applied through any amplifier-load system for driving the load, the reference load correction signal on the recording will provide full correction to the load, just as in the case where the reference load correction system of the invention is physically connected in the amplifier-load line. With the reference load correction circuit thus used as a processor of sound, the speaker correction signal may be added in the commercial recording of records, sound tapes, video tapes, motion picture sound tracks, and the like. In this case, the processed correction signal is actually stored in the recording. Such an arrangement is illustrated in Figure 6, wherein the recording stage, generally designated 290, is shown connected between the output terminal 14 of reference load correction system 10 and the input terminal 16 of power amplifier 18.

Claims

1. A method of correcting for variations in a load (20) having inductive reactance and characteristics resembling those of capacitive reactance, which load is driven by a power amplifier (18) that has an input circuit line through which it

receives an input program signal having a variable voltage waveform, said method comprising the steps of:

producing a first correction signal by means of a separate first channel (118c);

producing a second correction signal by means of a separate second channel (120c);

characterized by

inverting said input program signal to provide an inverted program signal;

mixing said first and second correction signals with said inverted program signal and applying such mixture of signals to said power amplifier (18);

said input program signal being applied through said separate first high frequency channel (118c) to a high pass filter means (200) to produce said first high frequency correction signal which has a phase lead and frequency response that increase with increasing frequencies so as to compensate for inductive reactance of said load (20);

said input program signal being applied through said separate second low frequency channel (120c) to a low pass filter means (210) to produce said second low frequency correction signal which has a phase lag and frequency response that increase with decreasing frequencies so as to compensate for characteristics of said load which resemble capacitive reactance;

said input program signal being inverted to deliver the high frequency harmonics and transients present in the input program signal out to ear via the high frequency correction signal before the onset of midrange frequencies present in the input program signal.

2. The method of claim 1, wherein said program signal that is mixed with said high and low frequency correction signals is first passed through a separate midrange frequency channel (220) having midrange band pass filter means (224) therein.
3. The method of claim 2, wherein the signal amplitude is relatively adjustable between said high frequency channel (118c) and said midrange frequency channel (220).
4. The method of claim 2, wherein the signal amplitude of each of said high and low frequency channels (118c, 120c) is adjustable relative to the signal amplitude of said midrange channel (220).
5. The method of claim 3, wherein the amplitude of said high frequency channel (118d) is auto-



type of operational amplifier that has proved to be satisfactory in the present invention is a 4558 dual operational amplifier such as that currently available as a standard chip from Motorola, National and Raytheon. This is a high performance operational amplifier with low noise and fast response characteristics. Also the TLO 62 operational amplifier from Texas Instruments has been found satisfactory. Also by way of example only and not of limitation, a part that has been found satisfactory for the variable resistor FET 242 of Figure 5 is a Siliconics J 230FET.

The various circuit component values shown in the circuit diagrams of Figures 1-5 have been found to produce satisfactory results in experimental prototypes, but these also are given by way of example only and not of limitation.

Although the present invention has been shown and described herein primarily in an arrangement in which it provides reference load correction signal directly into the amplification line which drives a load such as a speaker system, the invention may also be employed in the processing of recorded program information. Thus, the original program, whether it be live or prerecorded, is applied to an amplification line that embodies the reference load correction system of the invention, and the resulting output program that includes the correction voltage introduced by the reference load system of the invention is recorded (or re-recorded if the program came from a prior recording). Then, when the processed recording with the reference load correction signal thereon is applied through any amplifier-load system for driving the load, the reference load correction signal on the recording will provide full correction to the load, just as in the case where the reference load correction system of the invention is physically connected in the amplifier-load line. With the reference load correction circuit thus used as a processor of sound, the speaker correction signal may be added in the commercial recording of records, sound tapes, video tapes, motion picture sound tracks, and the like. In this case, the processed correction signal is actually stored in the recording. Such an arrangement is illustrated in Figure 6, wherein the recording stage, generally designated 290, is shown connected between the output terminal 14 of reference load correction system 10 and the input terminal 16 of power amplifier 18.

Claims

1. A method of correcting for variations in a load (20) having inductive reactance and characteristics resembling those of capacitive reactance, which load is driven by a power amplifier (18) that has an input circuit line through which it

receives an input program signal having a variable voltage waveform, said method comprising the steps of:

producing a first correction signal by means of a separate first channel (118c);
producing a second correction signal by means of a separate second channel (120c);
characterized by

inverting said input program signal to provide an inverted program signal;

mixing said first and second correction signals with said inverted program signal and applying such mixture of signals to said power amplifier (18);

said input program signal being applied through said separate first high frequency channel (118c) to a high pass filter means (200) to produce said first high frequency correction signal which has a phase lead and frequency response that increase with increasing frequencies so as to compensate for inductive reactance of said load (20);

said input program signal being applied through said separate second low frequency channel (120c) to a low pass filter means (210) to produce said second low frequency correction signal which has a phase lag and frequency response that increase with decreasing frequencies so as to compensate for characteristics of said load which resemble capacitive reactance;

said input program signal being inverted to deliver the high frequency harmonics and transients present in the input program signal out to ear via the high frequency correction signal before the onset of midrange frequencies present in the input program signal.

2. The method of claim 1, wherein said program signal that is mixed with said high and low frequency correction signals is first passed through a separate midrange frequency channel (220) having midrange band pass filter means (224) therein.

3. The method of claim 2, wherein the signal amplitude is relatively adjustable between said high frequency channel (118c) and said midrange frequency channel (220).

4. The method of claim 2, wherein the signal amplitude of each of said high and low frequency channels (118c, 120c) is adjustable relative to the signal amplitude of said midrange channel (220).

5. The method of claim 3, wherein the amplitude of said high frequency channel (118d) is auto-



matically adjustable relative to the amplitude of said midrange channel (244).

6. The method of claim 5, wherein said automatic adjustability is accomplished by developing a negative DC analog of the high frequency channel signal amplitude, a positive DC analog of the midrange channel signal amplitude, summing such negative and positive DC analogs, and using such sum to adjust circuit component means (242) in said high frequency channel (118d). 5
7. The method of claim 6, wherein such adjustable circuit component means (242) comprises FET means arranged in a resistor configuration. 10 15
8. The method of claim 2, wherein the signal of one of said high frequency and midrange frequency channels (118d, 244) is deliberately phase shifted to give the high frequency channel signal a substantial phase lead relative to the midrange frequency channel signal. 20
9. The method of claim 8, wherein said phase lead is approximately 180°. 25
10. A load correction system for a power amplifier (18) having a program input adapted for connection to a source of a program voltage signal that is variable as to waveform, and having an output connected to a driven load (20) having inductive reactance and characteristics resembling those of capacitive reactance, comprising: 30 35
 - a separate first channel circuit (118c) for producing a first correction signal,
 - a separate second channel circuit (120c) for producing a second correction signal, and 40
 - a separate program frequency channel circuit (220),
 - each of said channel circuits (118c, 120c, 220) having an input that is electrically connected to said program signal source (12g) and having an output, 45
 - said outputs of said channel circuits (118c, 120c, 220) being electrically connected to mixing circuit means (148c, 150c, 154c) so as to mix said correction signal components and said program signal, said mixing circuit means having an output (14g) that is electrically connected to said power amplifier input for applying such mixture of signals to said power amplifier (18); 50 55
 - characterized in
 - that said separate first channel circuit is a high frequency channel circuit (118c) contain-

ing a high pass filter (200) for producing said first correction signal which is a high frequency correction signal having a phase lead and frequency response that increase with increasing frequencies so as to compensate for inductive reactance of said load (20), and

that said separate second channel circuit is a low frequency channel circuit (120c) containing a low pass filter (210) for producing said second correction signal which is a low frequency correction signal having a phase lag and frequency response that increase with decreasing frequencies so as to compensate for characteristics of said load (20) which resemble capacitive reactance;

said separate program frequency channel circuit (244) inverting said program input signal to deliver the high frequency harmonics and transients present in the input program signal out to ear via the high frequency correction signal before the onset of midrange frequencies present in the input program signal.

11. The system of claim 10, 25
 - wherein said program frequency channel circuit comprises a midrange frequency channel circuit (220) having midrange band pass filter means (224) therein.
12. The system of claim 11, which comprises amplitude adjusting means (144c) in at least one of said high frequency and midrange frequency channel circuits (118c, 220), enabling relative adjustment between the amplitudes of the signals in said high frequency and midrange frequency channel circuits (118c, 220). 30 35
13. The system of claim 11, which comprises amplitude adjusting means (144c, 146c) in each of said high and low frequency channel circuits (118c, 120c), enabling the amplitudes of said signals in both of said high and low frequency channel circuits (118c, 120c) to be adjusted relative to the amplitude of the signal in said midrange channel circuit (220). 40 45
14. The system of claim 12, wherein said amplitude adjusting means (242) is automatic.
15. The system of claim 14, wherein said automatic amplitude adjusting means comprises first detector means (256) in said midrange channel circuit (244) providing a positive DC analog of the midrange frequency signal amplitude, 50 55
 - second detector means (274) in said high frequency channel circuit (118d) providing a negative DC analog of the high frequency sig-



nal amplitude,

summing circuit means (260, 278, 264) connected to the outputs of said first and second detector means (256, 274) for summing said DC outputs, and gain adjusting means (242) in said high frequency channel circuit (118d) operatively connected to the output (280) of said summing circuit means for automatically adjusting the gain of said high frequency channel circuit (118d) in response to variations in the sum of said positive and negative DC voltages.

16. The system of claim 15, wherein said gain adjusting means (242) comprises FET means arranged in a resistor configuration.

17. The system of claim 11, which comprises further phase shifting means (248) in one of said high frequency and midrange frequency channel circuits (118d, 240) for providing the signal in said high frequency channel circuit (118d) with a substantial phase lead relative to the signal in said midrange frequency channel circuit (240).

18. The system of claim 17, wherein said further phase shifting means comprises inverter circuit means (248) in one of said midrange frequency and high frequency channel circuits (118d, 244).

19. The system of claim 18, wherein said inverter circuit means (248) is in said midrange channel circuit (244).

Revendications

1. Procédé de correction de variations dans une charge (20) ayant une réactance inductive et des caractéristiques ressemblant à celles d'une réactance capacitive, cette charge étant excitée par un amplificateur de puissance (18) qui a une ligne de circuit d'entrée par l'intermédiaire de laquelle il reçoit un signal de programme d'entrée ayant une forme d'onde de tension variable, ledit procédé comprenant les étapes de:

production d'un premier signal de correction à l'aide d'un premier canal séparé (118c);

production d'un second signal de correction à l'aide d'un second canal séparé (120c);

caractérisé par

une inversion dudit signal de programme d'entrée pour réaliser un signal de programme inversé;

un mélange desdits premier et second signaux de correction avec ledit signal de pro-

gramme inversé et l'application de ce mélange de signaux audit amplificateur de puissance (18);

ledit signal de programme d'entrée étant appliqué via ledit premier canal haute fréquence séparé (118c) à un moyen formant filtre passe-haut (200) pour produire ledit premier signal de correction de haute fréquence qui a une avance de phase et une réponse en fréquence qui augmentent à mesure qu'augmentent les fréquences de façon à compenser la réactance inductive de ladite charge (20);

ledit signal de programme d'entrée étant appliqué via ledit second canal basse fréquence séparé (120c) à un moyen formant filtre passe-bas (210) pour produire ledit second signal de correction de basse fréquence qui a un retard de phase et une réponse en fréquence qui augmentent à mesure que diminuent les fréquences de façon à compenser les caractéristiques de ladite charge qui ressemblent à une réactance capacitive;

ledit signal de programme d'entrée étant inversé pour transmettre les harmoniques et transitoires haute fréquence présentes dans le signal de programme d'entrée via le signal de correction de haute fréquence avant l'apparition des moyennes fréquences présentes dans le signal de programme d'entrée.

2. Procédé selon la revendication 1, dans lequel ledit signal de programme mélangé avec lesdits signaux de correction de haute et de basse fréquence est d'abord amené à passer par un canal moyenne fréquence séparé (220) dans lequel se trouve un moyen formant filtre passe-bande (224) pour moyennes fréquences.

3. Procédé selon la revendication 2, dans lequel l'amplitude du signal peut subir un réglage relatif entre ledit canal haute fréquence (118c) et ledit canal moyenne fréquence (220).

4. Procédé selon la revendication 2, dans lequel l'amplitude du signal de chacun des canaux haute et basse fréquence (118c, 120c) est réglable par rapport à l'amplitude du signal dudit canal moyenne fréquence (220).

5. Procédé selon la revendication 3, dans lequel l'amplitude dudit canal haute fréquence (118c) est réglable d'une façon automatique par rapport à l'amplitude dudit canal moyenne fréquence (244).

6. Procédé selon la revendication 5, dans lequel ladite possibilité de réglage automatique est obtenue en élaborant un équivalent à courant



- continu négatif de l'amplitude du signal du canal haute fréquence, un équivalent à courant continu positif de l'amplitude du signal du canal moyenne fréquence, en additionnant ces équivalents à courant continu négatif et positif et en utilisant cette somme pour régler un moyen formant composant (242) de circuit dans ledit canal haute fréquence (118d).
7. Procédé selon la revendication 6, dans lequel ce moyen réglable formant composant (242) de circuit comporte un moyen formant TEC agencé sous la forme d'une résistance.
8. Procédé selon la revendication 2, dans lequel le signal d'un desdits canaux haute fréquence et moyenne fréquence (118d, 244) est délibérément déphasé pour donner au signal du canal haute fréquence une avance de phase notable par rapport au signal du canal moyenne fréquence.
9. Procédé selon la revendication 8, dans lequel ladite avance de phase est d'environ 180°.
10. Système de correction de charge pour un amplificateur de puissance (18) ayant une entrée de programme conçue pour être reliée à une source d'un signal de tension de programme qui a une forme d'onde variable, et ayant une sortie reliée à une charge excitée (20) ayant une réactance inductive et des caractéristiques ressemblant à celles d'une réactance capacitive, comprenant :
- un circuit de premier canal séparé (118c) pour produire un premier signal de correction,
 - un circuit de second canal séparé (120c) pour produire un second signal de correction, et
 - un circuit de canal séparé (220) de fréquence de programme,
- chacun desdits circuits de canaux (118c, 120c, 220) ayant une entrée reliée électriquement à ladite source (12g) de signal de programme et ayant une sortie,
- lesdites sorties desdits circuits de canaux (118c, 120c, 220) étant reliées électriquement à des moyens formant circuit mélangeur (148c, 150c, 154c) de façon à mélanger lesdits signaux de correction et ledit signal de programme, lesdits moyens formant circuit mélangeur ayant une sortie (14g) reliée électriquement à ladite entrée de l'amplificateur de puissance pour appliquer ce mélange de signaux audit amplificateur de puissance (18);
- caractérisé**
- en ce que ledit circuit de premier canal séparé est un circuit de canal haute fréquence (118c) contenant un filtre passe-haut (200) pour produire ledit premier signal de correction qui est un signal de correction de haute fréquence ayant une avance de phase et une réponse en fréquence qui augmentent à mesure qu'augmentent les fréquences de façon à compenser la réactance inductive de ladite charge (20), et
- en ce que ledit circuit de second canal séparé est un circuit de canal basse fréquence (120c) contenant un filtre passe-bas (210) pour produire ledit second signal de correction qui est un signal de correction de basse fréquence ayant un retard de phase et une réponse en fréquence qui augmentent à mesure que diminuent les fréquences de façon à compenser les caractéristiques de ladite charge (20) qui ressemblent à une réactance capacitive;
- ledit circuit de canal séparé (244) de fréquence de programme inversant ledit signal d'entrée de programme pour transmettre les harmoniques et transitoires haute fréquence présentes dans le signal de programme d'entrée via le signal de correction de haute fréquence avant l'apparition des moyennes fréquences présentes dans le signal de programme d'entrée.
11. Système selon la revendication 10, dans lequel ledit circuit de canal de fréquence de programme comporte un circuit de canal de moyenne fréquence (220) contenant un moyen formant filtre passe-bande (224) pour moyennes fréquences.
12. Système selon la revendication 11, comportant un moyen de réglage d'amplitude (144c) dans au moins un desdits circuits de canaux haute fréquence et moyenne fréquence (118c, 220), permettant un réglage relatif entre les amplitudes des signaux dans lesdits circuits de canaux haute fréquence et moyenne fréquence (118c, 220).
13. Système selon la revendication 11, comportant un moyen de réglage d'amplitude (144c, 146c) dans chacun desdits circuits de canaux haute et basse fréquence (118c, 120c), permettant aux amplitudes desdits signaux présents dans lesdits deux circuits de canaux haute et basse fréquence (118c, 120c) d'être réglées par rapport à l'amplitude du signal présent dans ledit circuit de canal moyenne fréquence (220).
14. Système selon la revendication 12, dans lequel ledit moyen de réglage d'amplitude (242) est automatique.



15. Système selon la revendication 14, dans lequel ledit moyen de réglage automatique d'amplitude comporte un premier moyen de détection (256) présent dans ledit circuit de canal moyenne fréquence (244) réalisant un équivalent à courant continu positif de l'amplitude du signal moyenne fréquence,

un second moyen de détection (274) présent dans ledit circuit de canal haute fréquence (118d) réalisant un équivalent à courant continu négatif de l'amplitude du signal haute fréquence,

des moyens formant circuit sommateur (260, 278, 264) reliés aux sorties desdits premier et second moyens de détection (256, 274) pour additionner lesdites sorties de courant continu, et un moyen de réglage de gain (242) présent dans ledit circuit de canal haute fréquence (118d) coopérant avec la sortie (280) desdits moyens formant circuit sommateur pour régler automatiquement le gain dudit circuit de canal haute fréquence (118d) en réponse aux variations de la somme desdites tensions continues positive et négative.

16. Système selon la revendication 15, dans lequel ledit moyen de réglage de gain (242) comporte un moyen formant TEC agencé sous la forme d'une résistance.

17. Système selon la revendication 11, comportant un autre moyen de déphasage (248) présent dans un desdits circuits de canaux haute fréquence et moyenne fréquence (118d, 240) pour donner à un signal présent dans ledit circuit de canal haute fréquence (118d) une avance de phase notable par rapport au signal présent dans ledit circuit de canal moyenne fréquence (240).

18. Système selon la revendication 17, dans lequel ledit autre moyen de déphasage comporte un moyen formant circuit inverseur (248) dans l'un desdits circuits de canaux moyenne fréquence et haute fréquence (118d, 244).

19. Système selon la revendication 18, dans lequel ledit moyen formant circuit inverseur (248) est dans ledit circuit de canal moyenne fréquence (244).

Patentansprüche

1. Verfahren zum Korrigieren von Schwankungen in einer Last (20), die einen induktiven Blindwiderstand sowie einem kapazitiven Blindwiderstand ähnliche Eigenschaften aufweist und die von einem Leistungsverstärker (18) angesteuert

ist, der eine Eingangsleitung aufweist, über die er ein Eingangs-Programmsignal mit einer Wellenform variabler Spannung empfängt, wobei das Verfahren folgende Schritte umfaßt:

Erzeugen eines ersten Korrektursignals mittels eines separaten ersten Kanals (118c), und

Erzeugen eines zweiten Korrektursignals mittels eines separaten zweiten Kanals (120c), gekennzeichnet durch

Invertieren des Eingangs-Programmsignals zur Erzeugung eines invertierten Programmsignals,

Mischen des ersten und des zweiten Korrektursignals mit dem invertierten Programmsignal und Beaufschlagen des Leistungsverstärkers (18) mit diesem Signalgemisch,

wobei das Eingangs-Programmsignal über den separaten ersten Hochfrequenzkanal (118c) einer Hochpaß-Filtereinrichtung (200) zugeführt wird, um das erste hochfrequente Korrektursignal zu erzeugen, das eine mit steigender Frequenz zunehmende Phasenvorstellung und Frequenzabhängigkeit aufweist derart, daß induktiver Blindwiderstand der Last (20) kompensiert wird,

wobei das Eingangs-Programmsignal über den separaten zweiten Niederfrequenzkanal (120c) einer Tiefpaß-Filtereinrichtung (210) zugeführt wird, um ein zweites niederfrequentes Korrektursignal zu erzeugen, das eine mit abnehmender Frequenz zunehmende Phasennaheilung und Frequenzabhängigkeit aufweist derart, daß einem kapazitiven Blindwiderstand ähnliche Eigenschaften der Last kompensiert werden, und

wobei das Eingangs-Programmsignal invertiert wird, um die in ihm vorhandenen hochfrequenten Oberwellen und Einschwingvorgänge über das hochfrequente Korrektursignal dem Ohr zuzuführen, bevor in dem Eingangs-Programmsignal vorhandene mittlere Frequenzbereiche einsetzen.

2. Verfahren nach Anspruch 1, wobei das mit dem hoch- und niederfrequenten Korrektursignalen gemischte Programmsignal zunächst über einen getrennten Mittelfrequenzkanal (220) mit einer darin angeordneten Mittelbereichs-Bandpaßfiltereinrichtung (224) geleitet werden.

3. Verfahren nach Anspruch 2, wobei die Signalamplitude zwischen dem Hochfrequenzkanal (118c) und dem Mittelfrequenzkanal (220) relativ einstellbar ist.

4. Verfahren nach Anspruch 2, wobei die Signa-



lamplitude sowohl des Hochfrequenz- als auch des Niederfrequenzkanals (118c, 120c) relativ zur Signalamplitude des Mittelbereichskanals (220) einstellbar ist.

5. Verfahren nach Anspruch 3, wobei die Amplitude des Hochfrequenzkanals (118d) relativ zur Amplitude des Mittelbereichskanals (244) automatisch einstellbar ist.

6. Verfahren nach Anspruch 5, wobei die automatische Einstellbarkeit dadurch erreicht wird, daß ein negatives Gleichspannungsanalog der Signalamplitude auf dem Hochfrequenzkanal und ein positives Gleichspannungsanalog der Signalamplitude auf dem Mittelbereichskanal entwickelt, diese negativen und positiven Gleichspannungsanaloge summiert und die Summe zur Einstellung einer Schaltungsstufe (242) in dem Hochfrequenzkanal (118d) verwendet wird.

7. Verfahren nach Anspruch 6, wobei die einstellbare Schaltungsstufe (242) eine in Widerstandsschaltung angeordnete Feldeffekttransistor-Einrichtung umfaßt.

8. Verfahren nach Anspruch 2, wobei das Signal des Hochfrequenz- oder des Mittelfrequenzbereichskanals (118d, 244) absichtlich phasenverschoben wird, um dem Signal auf dem Hochfrequenzkanal eine wesentliche Phasenvoreilung gegenüber dem Signal auf dem Mittelfrequenzkanal zu erteilen.

9. Verfahren nach Anspruch 8, wobei die Phasenvoreilung etwa 180° beträgt.

10. Lastkorrektursystem für einen Leistungsverstärker (18) mit einem Programmeingang zum Anschluß an eine Quelle mit einem Programmspannungssignal variabler Wellenform und einem Ausgang, der mit einer angesteuerten Last (20) verbunden ist, die einen induktiven Blindwiderstand und einem kapazitiven Blindwiderstand ähnliche Eigenschaften aufweist, umfassend

eine separate erste Kanalstufe (118c) zur Erzeugung eines ersten Korrektursignals,

eine separate zweite Kanalstufe (120c) zur Erzeugung eines zweiten Korrektursignals, und

eine separate Programmfrequenz-Kanalstufe (220),

wobei jede der Kanalstufen (118c, 120c, 220) einen mit der Programmsignalquelle (12g) elektrisch verbundenen Eingang und einen Ausgang aufweist,

wobei die Ausgänge der Kanalstufen

(118c, 120c, 220) mit einer Mischstufe (148c, 150c, 154c) elektrisch verbunden sind, um die Korrektursignalkomponenten und das Programmsignal zu mischen, und wobei die Mischstufe einen mit dem Leistungsverstärker-eingang elektrisch verbundenen Ausgang (14g) aufweist, um das Signalgemisch dem Leistungsverstärker (18) zuzuführen,

dadurch gekennzeichnet,

daß die separate erste Kanalstufe eine Hochfrequenz-Kanalstufe (118c) mit einem Hochpaßfilter (200) zur Erzeugung des ersten Korrektursignals ist, bei dem es sich um ein hochfrequentes Korrektursignal handelt, das eine mit steigender Frequenz zunehmende Phasenvoreilung und Frequenzabhängigkeit aufweist derart, daß induktiver Blindwiderstand der Last (20) kompensiert wird, und

daß die separate zweite Kanalstufe eine Niederfrequenz-Kanalstufe (120c) mit einem Tiefpaßfilter (210) zur Erzeugung des zweiten Korrektursignals ist, bei dem es sich um ein niederfrequentes Korrektursignal handelt, das eine mit abnehmender Frequenz zunehmende Phasennacheilung und Frequenzabhängigkeit aufweist derart, daß einem kapazitiven Blindwiderstand ähnliche Eigenschaften der Last (20) kompensiert werden,

wobei die separate Programmfrequenz-Kanalstufe (244) das Eingangs-Programmsignal invertiert, um die in ihm vorhandenen hochfrequenten Oberwellen und Einschwingvorgänge über das hochfrequente Korrektursignal dem Ohr zuzuführen, bevor in dem Eingangs-Programmsignal vorhandene mittlere Frequenzbereiche einsetzen.

11. System nach Anspruch 10, wobei die Programmfrequenz-Kanalstufe eine Mittelfrequenz-Kanalstufe (220) mit einer darin vorgesehenen Mittelbereichs-Bandpaßfiltereinrichtung (224) umfaßt.

12. System nach Anspruch 11 mit in mindestens einer der Hochfrequenz- und Mittelfrequenz-Kanalstufen (118c, 120c) enthaltenen Amplituden-Einstelleinrichtung (144c), die eine relative Einstellung zwischen den Amplituden der Signale in der Hochfrequenz- und der Mittelfrequenz-Kanalstufe (118c, 220) gestattet.

13. System nach Anspruch 11 mit in jedem der Hochfrequenz- und Niederfrequenz-Kanalstufen (118c, 120c) enthaltenen Amplituden-Einstelleinrichtungen (144c, 146c), die eine Einstellung der Amplituden der Signale in sowohl der Hochfrequenz- als auch der Niederfrequenz-Kanalstufe (118c, 120c) relativ zur Amplitude



des Signals in der Mittelbereichs-Kanalstufe (220) gestatten.

14. System nach Anspruch 12, wobei die Amplituden-Einstelleinrichtung (242) automatisch ist. 5

15. System nach Anspruch 14, wobei die automatische Amplituden-Einstelleinrichtung umfaßt:
 - eine in der Mittelbereichs-Kanalstufe (244) enthaltene erste Detektoreinrichtung (256), die ein positives Gleichspannungsanalog der Amplitude des Mittelfrequenzsignals liefert, 10
 - eine in der Hochfrequenz-Kanalstufe (118d) angeordnete zweite Detektoreinrichtung (274), die ein negatives Gleichspannungsanalog der Amplitude des Hochfrequenzsignals liefert, 15
 - eine mit den Ausgängen der ersten und der zweiten Detektoreinrichtung (256, 274) verbundene Summierstufe (260, 278, 264) zum Aufsummieren der Gleichspannungsausgänge, und 20
 - eine in der Hochfrequenz-Kanalstufe (118d) vorhandene und mit dem Ausgang (280) der Summierstufe funktionsmäßig verbundene Verstärkungs-Einstelleinrichtung (242) zur automatischen Einstellung der Verstärkung der Hochfrequenz-Kanalstufe (118d) in Abhängigkeit von Änderungen in der Summe der positiven und negativen Gleichspannungen. 25 30

16. System nach Anspruch 15, wobei die Verstärkungs-Einstelleinrichtung (242) eine in Widerstandsschaltung angeordnete Feldeffekttransistor-Einrichtung umfaßt. 35

17. System nach Anspruch 11 mit einer in einer der Hochfrequenz- und Mittelfrequenz-Kanalstufen (118, 220) angeordnete weitere Phasenschiebeeinrichtung (248), die das Signal in der Hochfrequenz-Kanalstufe (118d) mit einer wesentlichen Phasenvoreilung gegenüber dem Signal in der Mittelfrequenz-Kanalstufe (240) versieht. 40 45

18. System nach Anspruch 17, wobei die weitere Phasenschiebeeinrichtung eine in einer der Mittel- und Hochfrequenz-Kanalstufen (118d, 244) angeordnete Inverterstufe (248) umfaßt. 50

19. System nach Anspruch 18, wobei die Inverterstufe (248) in der Mittelbereichs-Kanalstufe (244) liegt. 55