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(54) **POWER AMPLIFIER AND METHOD FOR SPLIT VOICE COIL TRANSDUCER OR SPEAKER**

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(57) **ABSTRACT**

The present invention relates to a method and power amplifier for providing an audio current comprising a first audio current and a second audio current to a split voice coil speaker having a first coil and a second coil located in a magnetic field. The first coil and the second coil are wired such that current can be provided to one coil without providing current to the other coil. The first audio current is provided to a first branch connected to the first coil. The second audio current is provided to a second branch connected to the second coil. The first audio current differs from the second audio current. The first audio current in the first coil and the second audio current in the second coil are oriented such that both the first audio current and the second audio current constructively contribute to a force provided to the voice coil.

Figure 1

Figure 2

Figure 3A   Figure 3B

Figure 4

Figure 5

Figure 6

Figure 7

Figure 9

Figure 8

SWITCHING FREQUENCY

R/8L

234

$a_b = 0.05$

236

$1/(T_1+T_2)$

$a_b = 0.1$
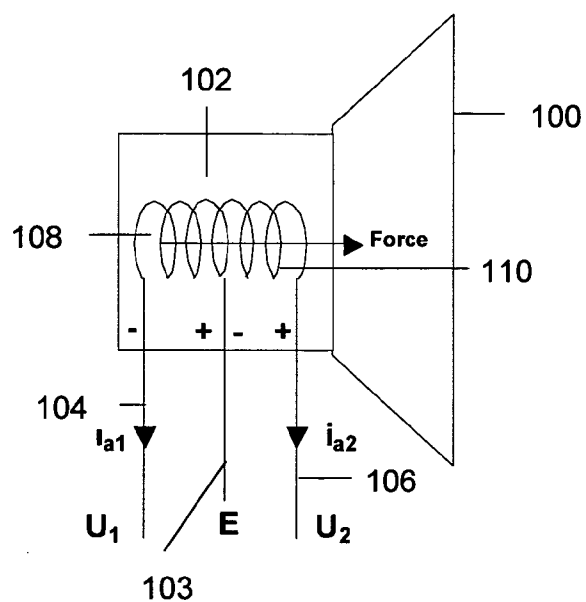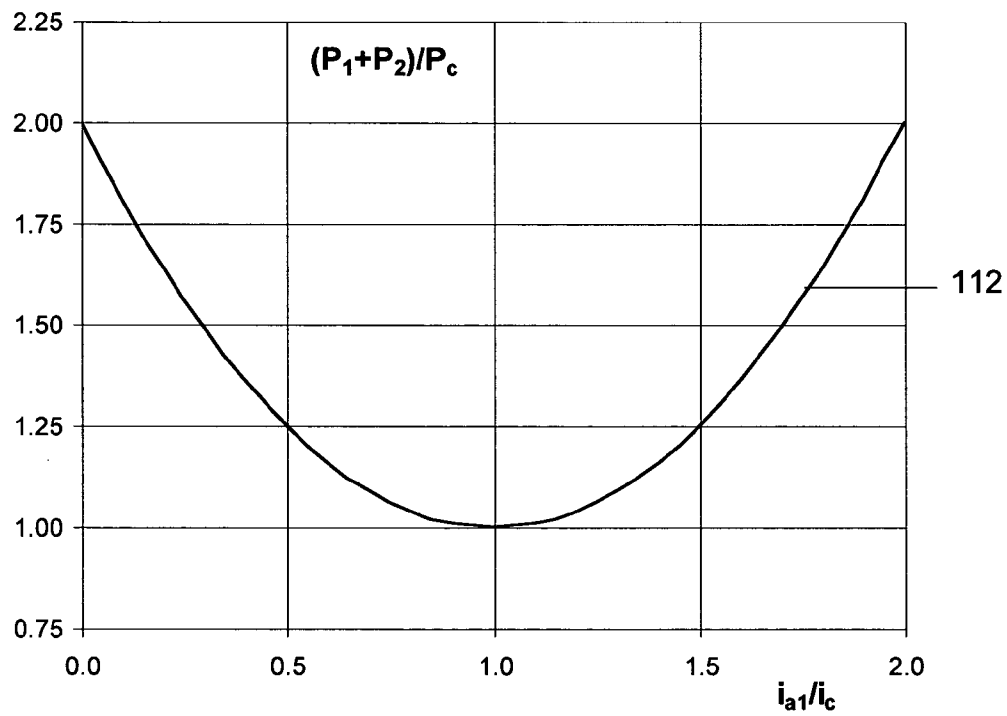
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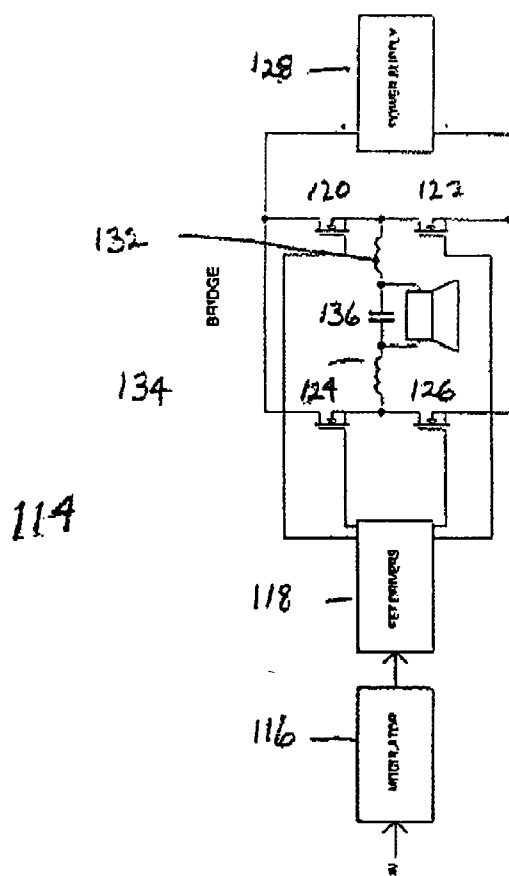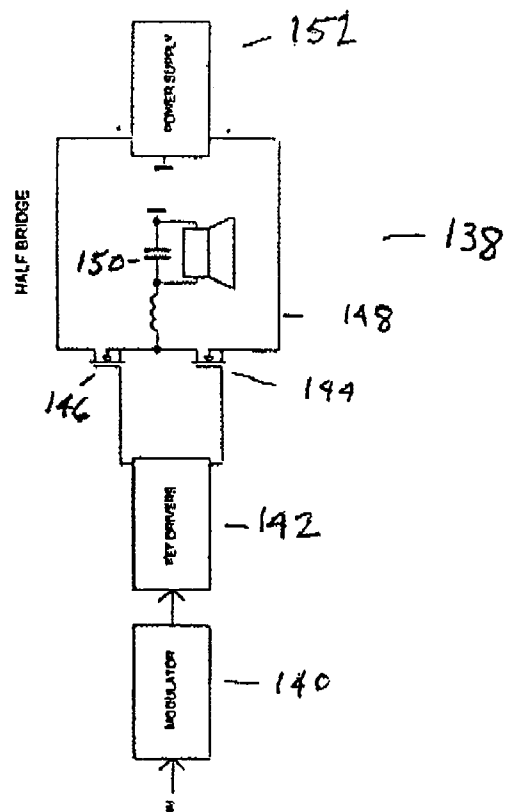|a|

Figure 10

DUTY CYCLE

$T_1/(T_1 + T_2)$

238

MODULATION
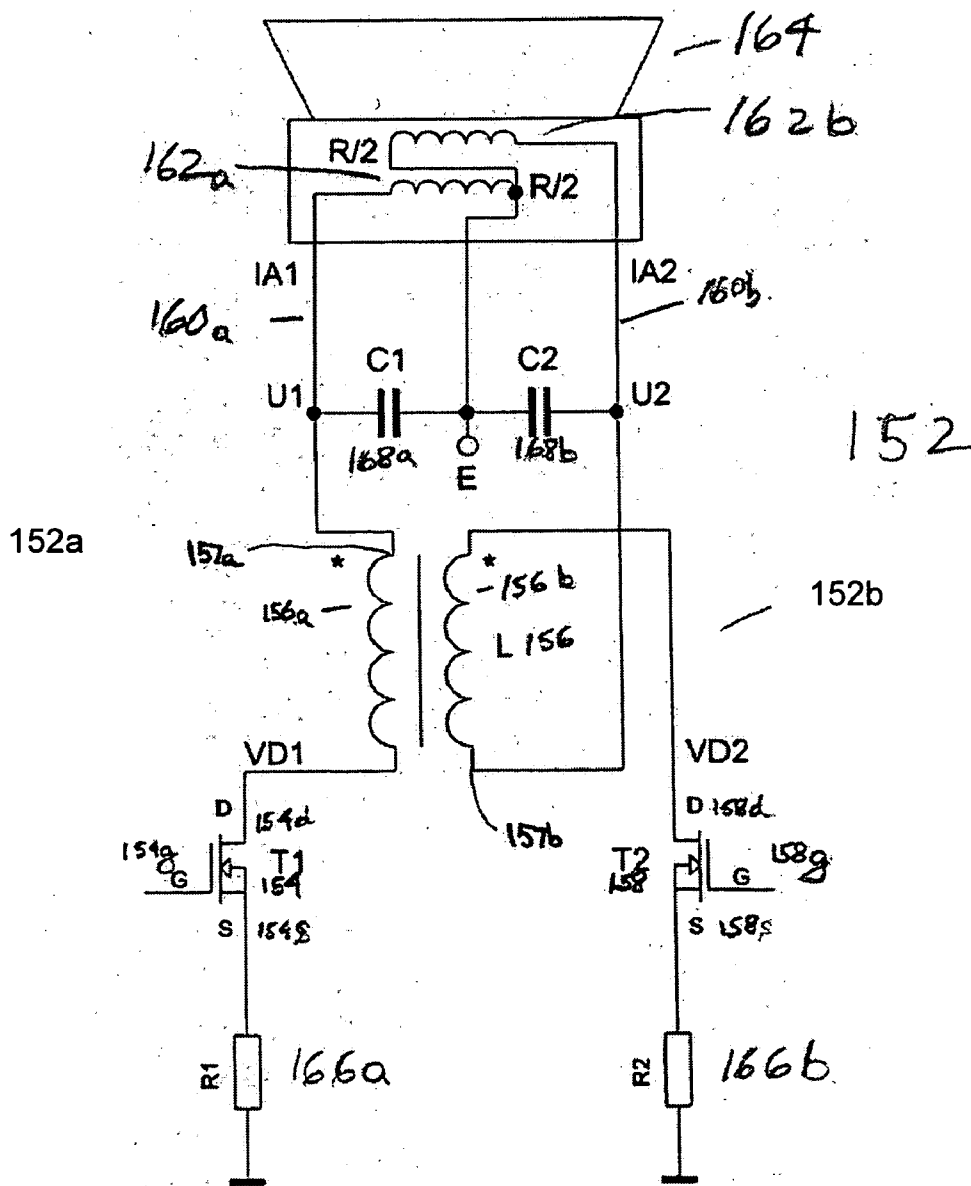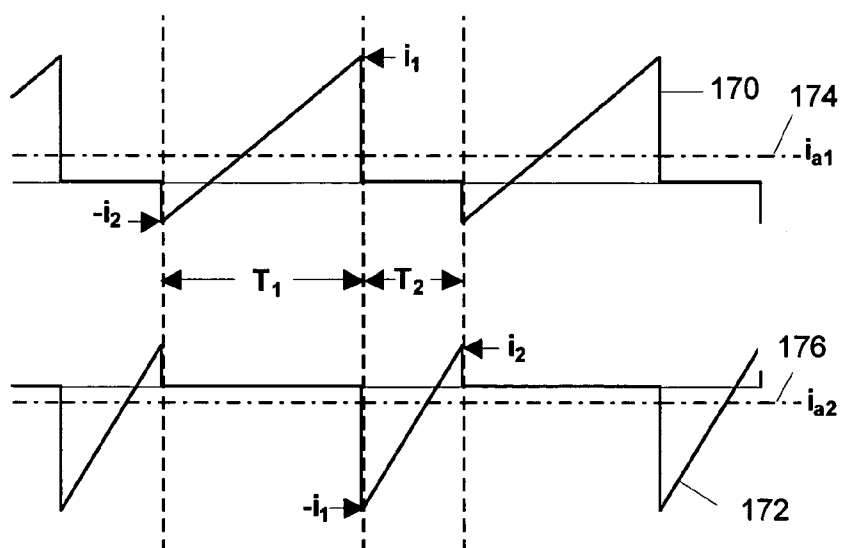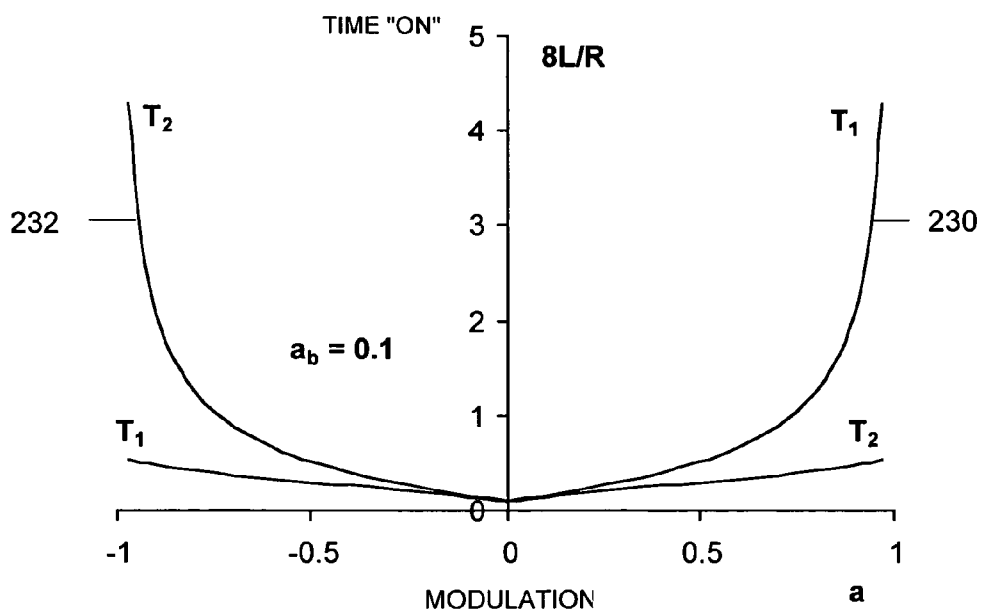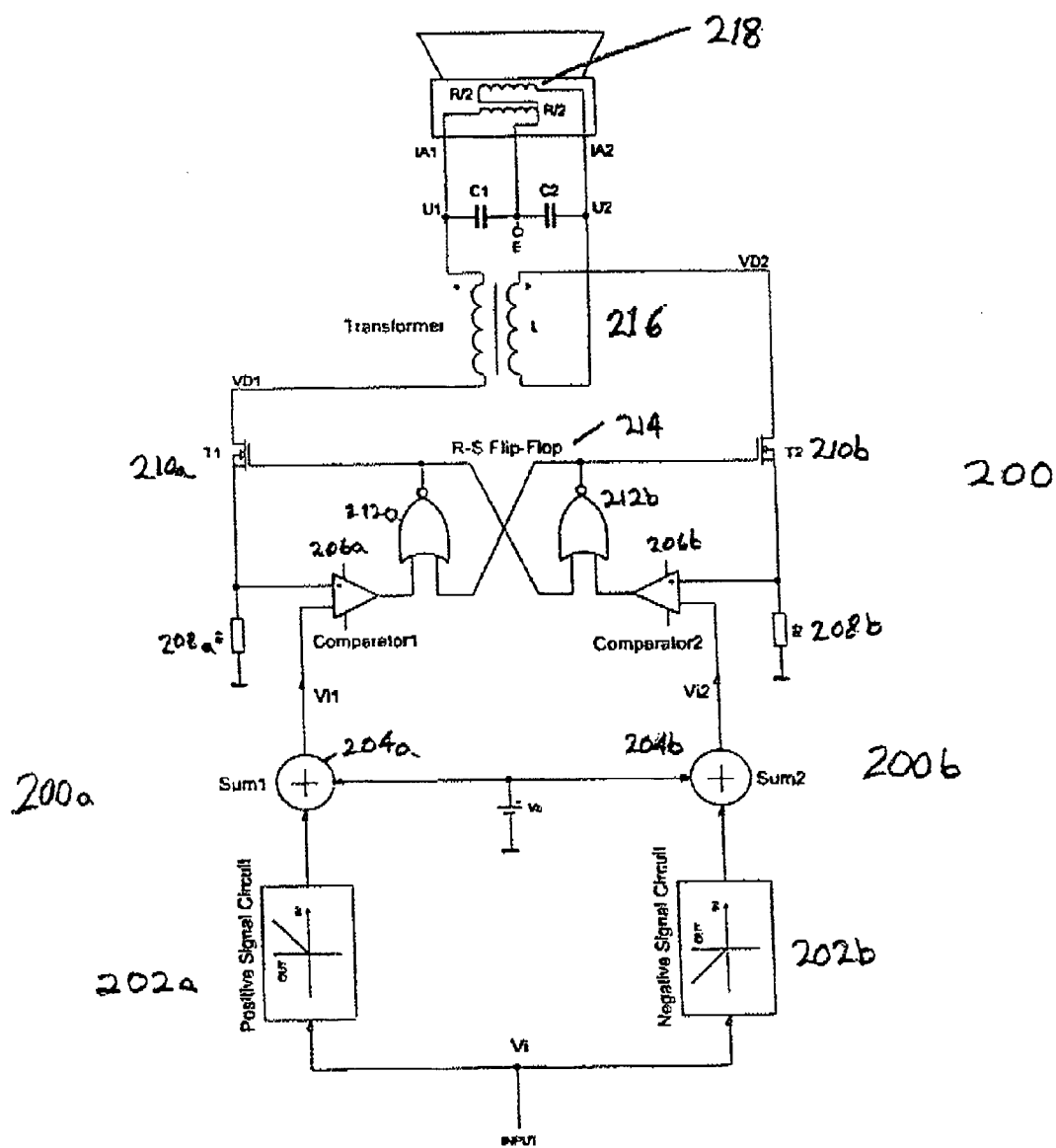
a

Figure 11

Figure 12

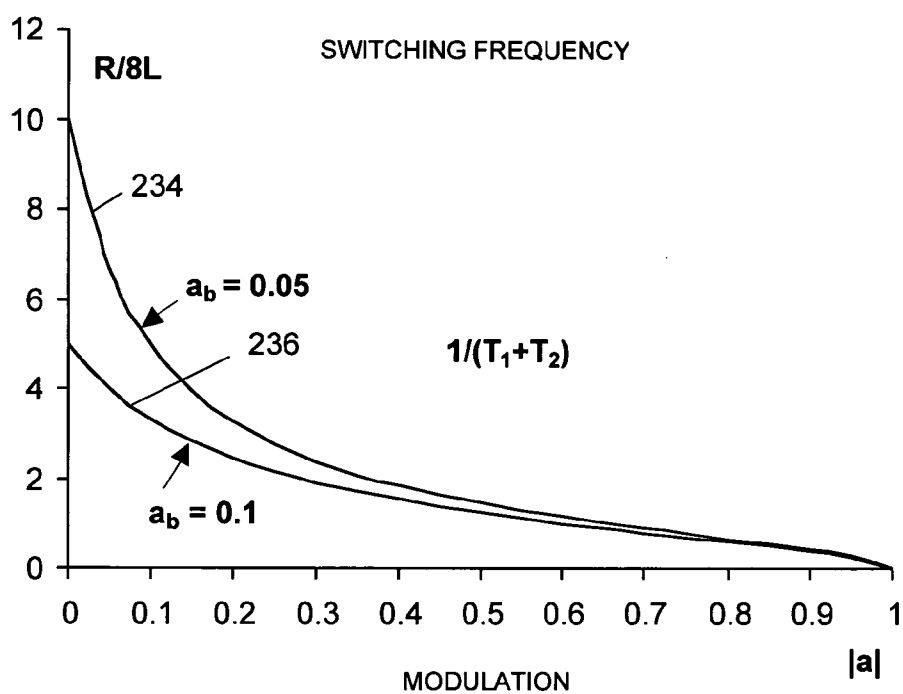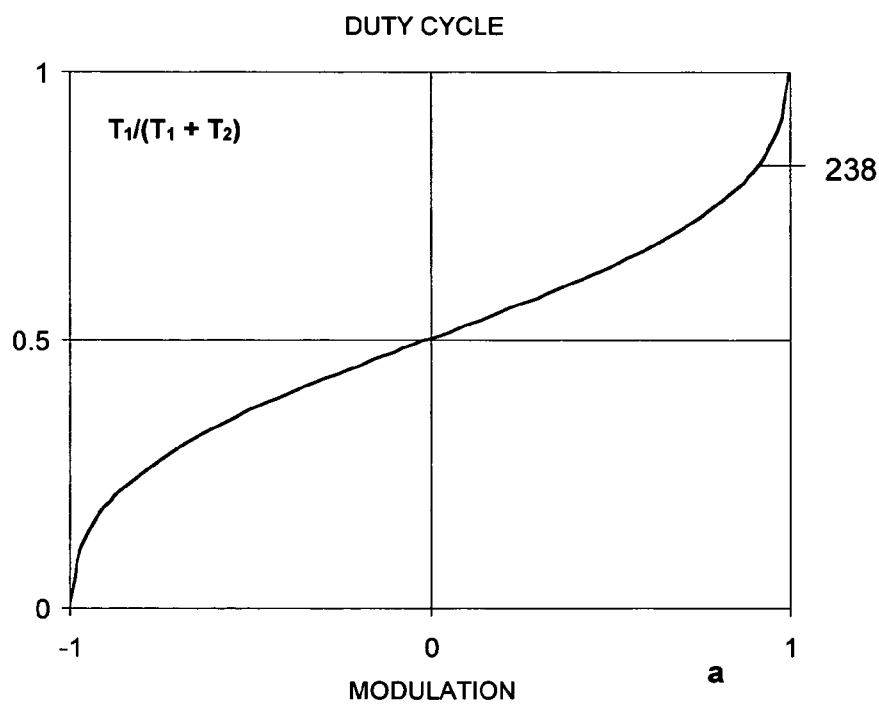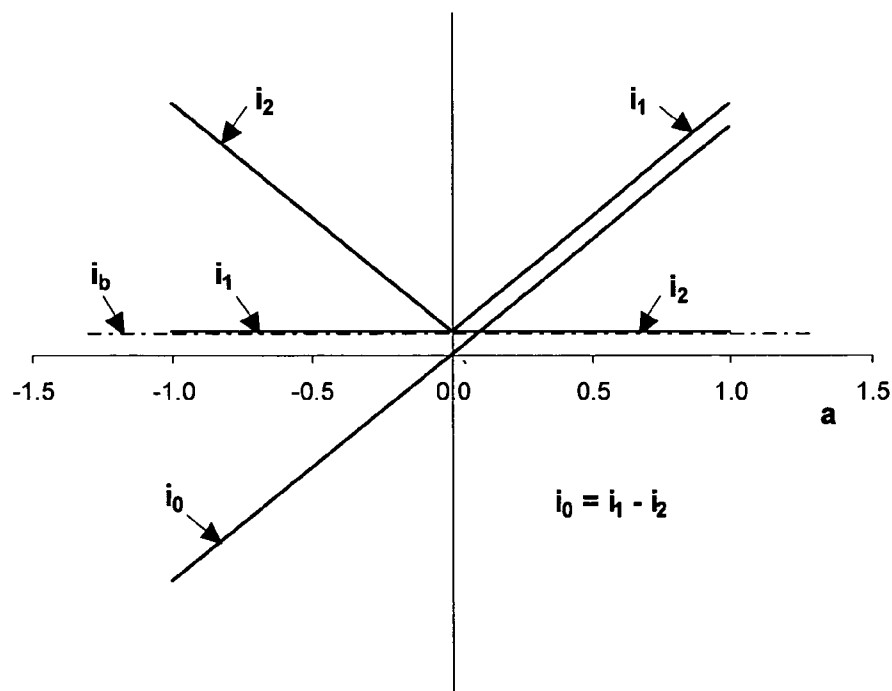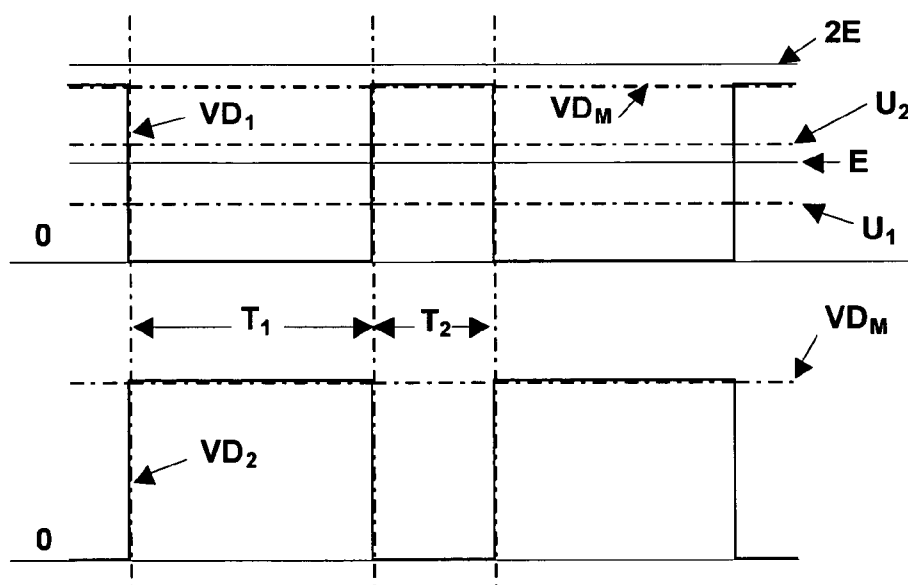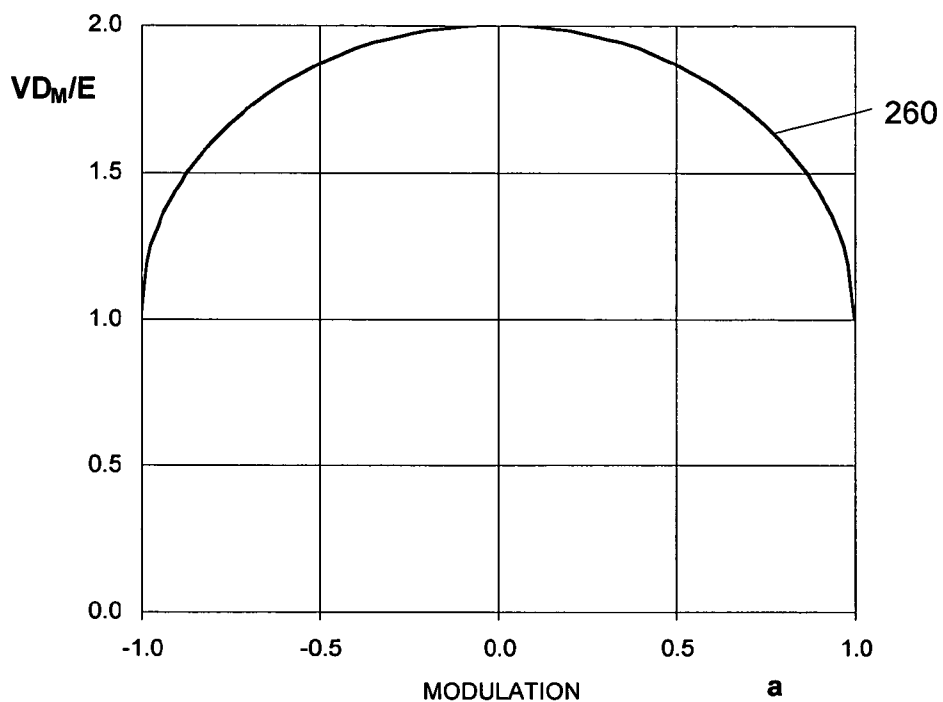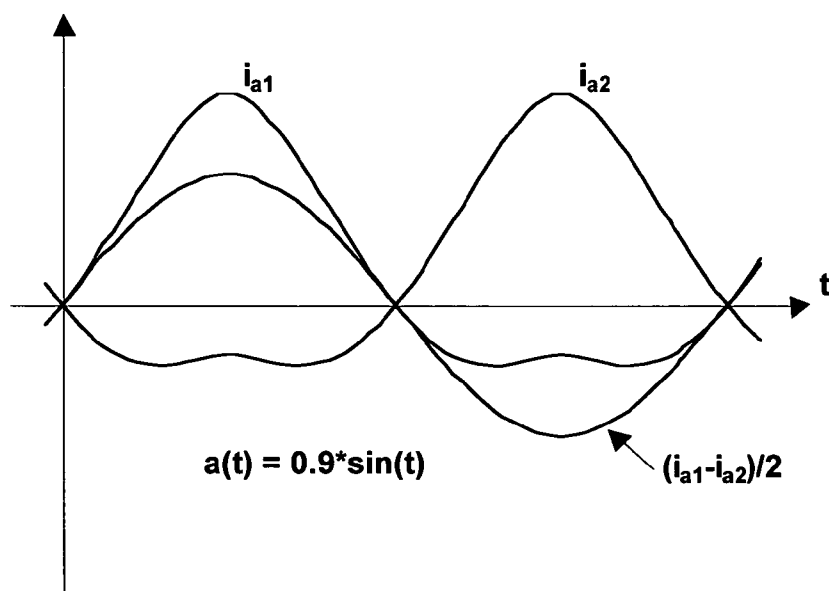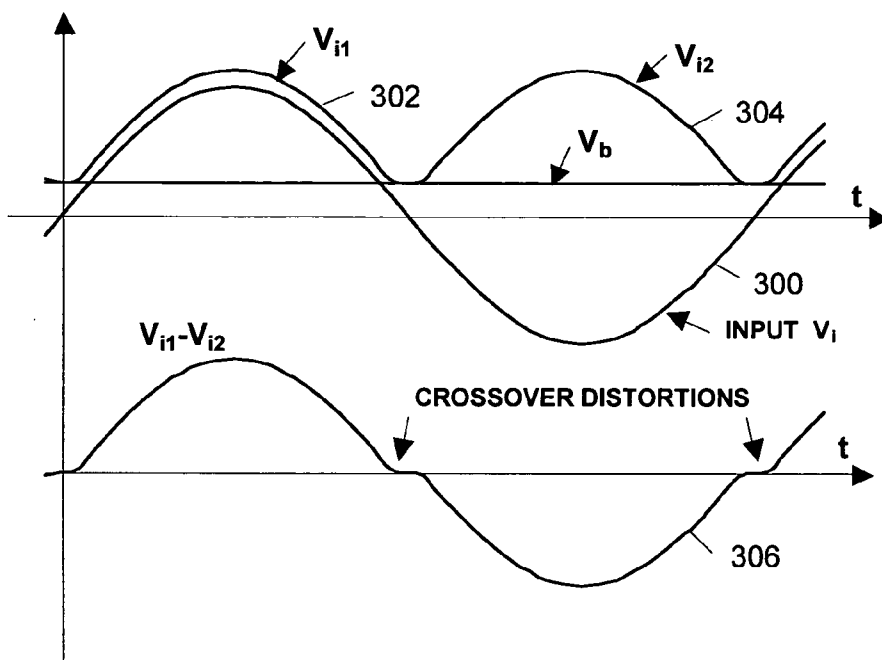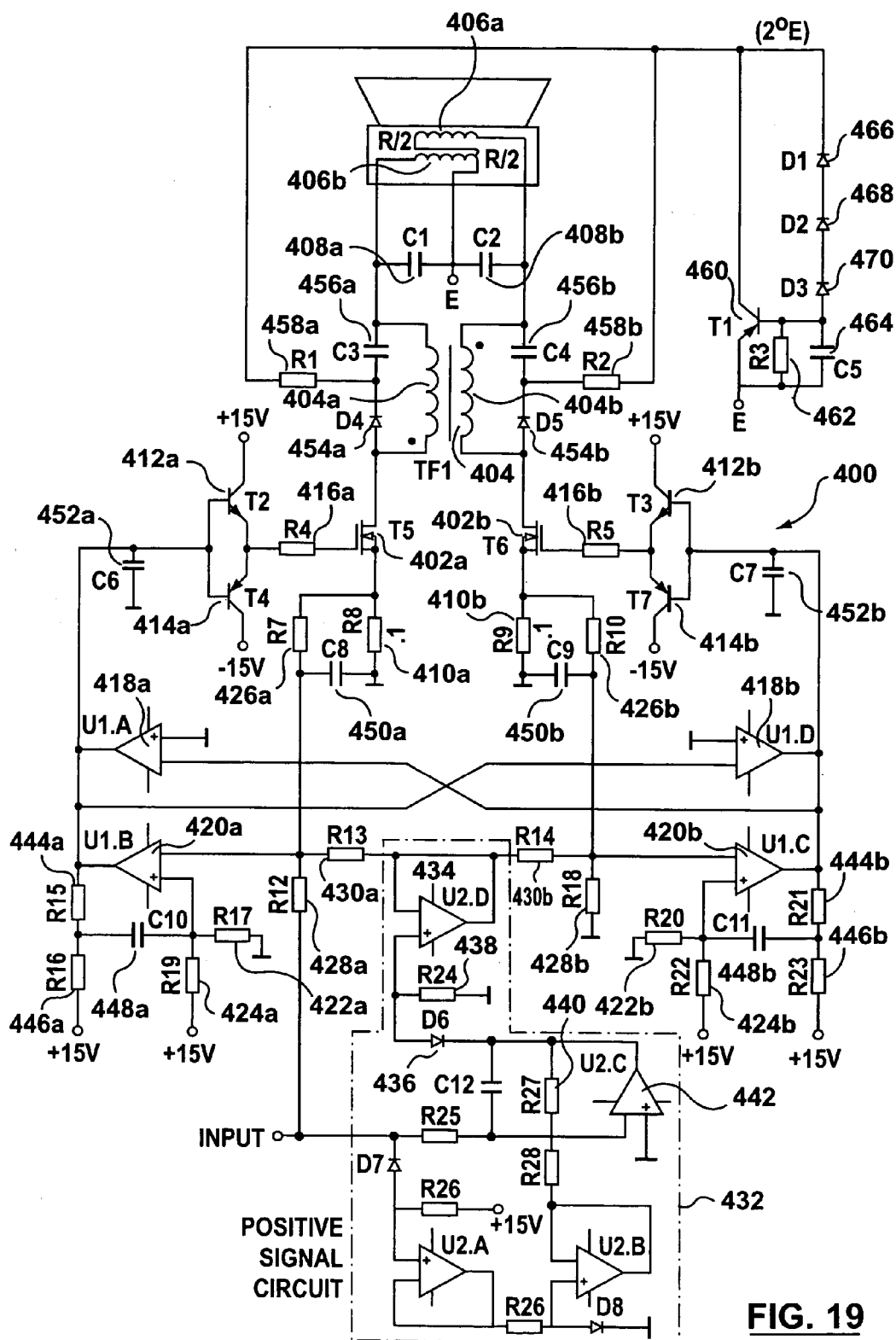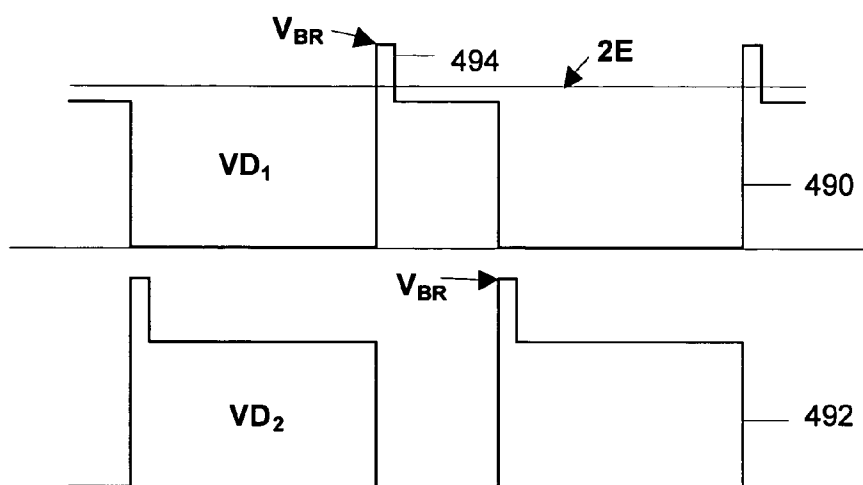
Figure 13

Figure 14

Figure 15

Figure 16

Figure 17

FIG. 18

FIG. 19

Figure 20

Figure 21

FIG. 22

TIME "ON"   5 ┐ **8L/R**

T₂

506

a₁ + a₂ = 1

a = a₁ - a₂

T₁

504

T₁   **T₂**

MODULATION

Figure 23

SWITCHING FREQUENC

R/8L

a₁ + a₂ = 1

a = a₁ - a₂

MODULATIOI
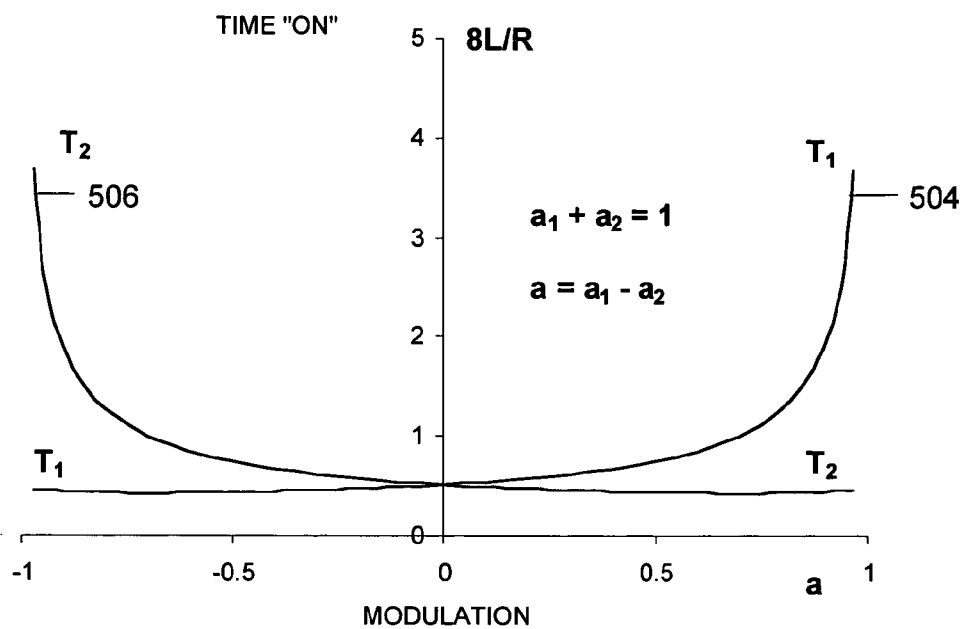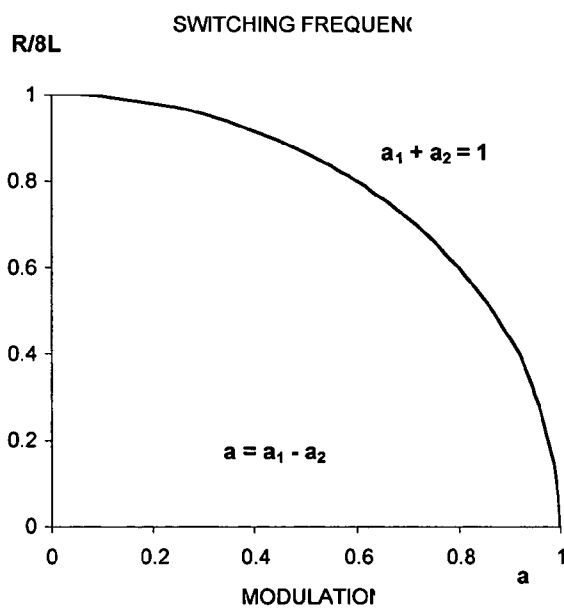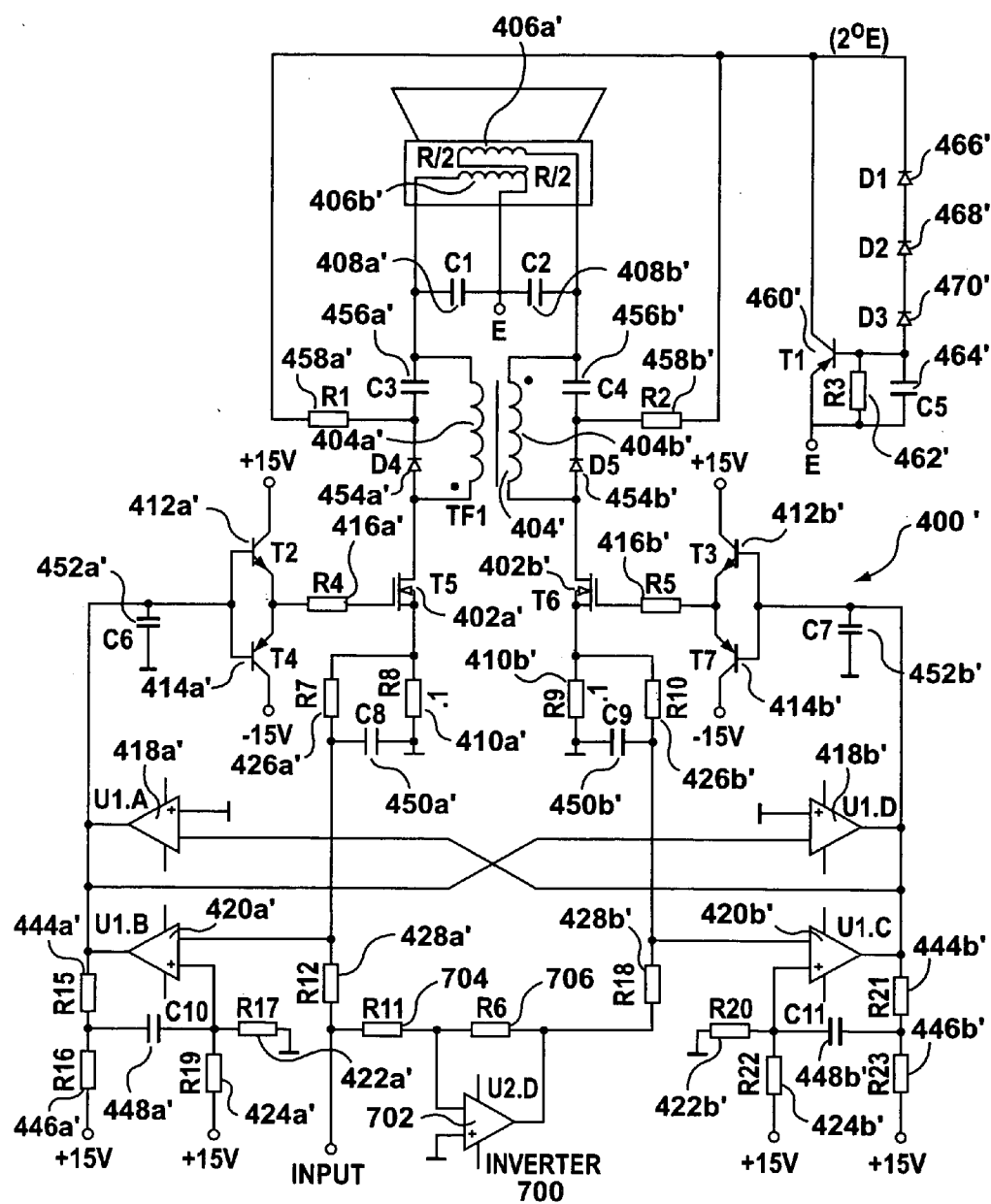
Figure 24

FIG. 25

POWER AMPLIFIER AND METHOD FOR SPLIT VOICE COIL TRANSDUCER OR SPEAKER

FIELD OF THE INVENTION

[0001] This invention relates generally to audio power amplifiers. More particularly, it relates to an audio power amplifier for a split voice coil transducer or speaker.

BACKGROUND OF THE INVENTION

[0002] As the cone of an operating speaker moves in both directions, the force generated by the speaker voice coil has to have an alternating character. This is achieved in conventional speakers by connecting an amplifier that supplies alternative or bi-directional current to the voice coil of a single coil speaker.

[0003] Power amplifiers are built using power devices such as vacuum tubes or transistors. These devices control the amount of current that flows from the power supply to the amplifier output based on the applied input signal. The vacuum tube or transistor modulates current intensity but cannot change current polarity. One reason for this is that the vacuum tube or transistor is connected to the power supply, which has constant polarity. The second reason comes from the fact that a vacuum tube or bipolar transistor can conduct current in one direction only. This is why two such power devices are required to build an amplifier. One power device is connected between a positive power supply and the amplifier output while the second power device is connected between a negative power supply and the amplifier output. The amplifier output acts as a subtracting node where signals from both devices meet.

[0004] The control of the power devices is simplified if they are connected in the same way to the circuit ground, particularly if their cathodes, emitters or sources are connected to the circuit ground. In this case, only one power supply, positive or negative, can be used, and an alternating current cannot be obtained (except when using a power audio transformer, which is costly and does not work at very low frequencies or DC). If this configuration is used to build a power amplifier without a power audio transformer, then some other mechanism must be provided to provide an alternating force in the speaker.

[0005] The split coil speaker provides such a mechanism for providing an alternating force in the speaker. Both coils together, work as subtracting device. That is, instead of subtracting currents one subtract forces with the same result for the speaker operation. In such speakers, each coil is placed in a magnetic field created by a magnet and is energized with an electric current. Each conducting coil then experiences a force created by the magnetic field of the magnet and the coil current. This force is used to drive a speaker cone to produce sound.

[0006] U.S. Pat. No. 2,959,640 (Schultz) and U.S. Pat. No. 4,130,725 (Nagel) disclose split coil transducers combined with class B amplifiers. Class B, push-pull or other amplifiers alternate the current between coils by delivering a current to one coil at a time. That is, while the current is supplied to one coil, no current is supplied to the other coil. However, for reasons that are discussed below, this reduces efficiency and increases heat dissipation in the transducer or speaker. It is possible to increase the density of the magnetic

flux in the transducer to reduce heat dissipation; however, this is expensive as it requires a much larger magnet.

[0007] It is very important for bifilar speaker systems to include power amplifiers capable of delivering high power at high efficiency, as reduced efficiency may impair cooling or increase cost due to the need to include large heat sinks. Accordingly, a high efficiency power amplifier for a split voice coil transducer is desirable.

SUMMARY OF THE INVENTION

[0008] In accordance with a first aspect of the present invention, there is provided a switching power amplifier for connection to a split voice coil of a speaker. The split voice coil has a first coil and a second coil located in a magnetic field, the first coil and the second coil being wired such that current can be provided to one coil without providing current to the other coil. The power amplifier comprises: a first branch for providing a first audio current to the first coil; a second branch for providing a second audio current to the second coil, wherein the first audio current differs from the second audio current; and a transformer for transferring energy between the first branch and the second branch such that both the first audio current and the second audio current constructively contribute to a force provided to the voice coil.

[0009] In accordance with a second aspect of the present invention, there is provided a method of providing an audio current comprising a first audio current and a second audio current to a split voice coil speaker having a first coil and a second coil located in a magnetic field. The first coil and the second coil are wired such that current can be provided to one coil without providing current to the other coil. The method comprises providing the first audio current to a first branch connected to the first coil; providing the second audio current to a second branch connected to the second coil, wherein the first audio current differs from the second audio current; orienting the first audio current in the first coil and the second audio current in the second coil such that both the first audio current and the second audio current constructively contribute to a force provided to the voice coil.

[0010] In accordance with a third aspect of the present invention, there is provided a speaker comprising a split voice coil having a first coil and a second coil located in a magnetic field, the first coil and the second coil being wired such that current can be provided to one coil without providing current to the other coil, and a power amplifier. The power amplifier comprises a first branch for providing a first audio current to the first coil; a second branch for providing a second audio current to the second coil, wherein the first audio current differs from the second audio current; and a transformer for transferring energy between the first branch and the second branch such that both the first audio current and the second audio current constructively contribute to a force provided to the voice coil.

BRIEF DESCRIPTION OF THE DRAWINGS

[0011] A detailed description of the preferred embodiments is provided herein below with reference to the following drawings in which:

[0012] FIG. 1, in a schematic view, illustrates a speaker with a split voice coil showing the polarity of applied voltages and currents;

[0013] FIG. 2 is a graph illustrating ratios of total power dissipated in the split voice coil of FIG. 1 under different conditions;

[0014] FIG. 3a is a schematic view of a full bridge class D power amplifier in accordance with the prior art;

[0015] FIG. 3b is a schematic view of a half bridge class D power amplifier in accordance with the prior art;

[0016] FIG. 4, in a simplified schematic diagram, illustrates a power amplifier in accordance with the present invention;

[0017] FIG. 5, in a graph, plots the power transistor currents in the amplifier of FIG. 4 as a function of time;

[0018] FIG. 6, in a graph, plots the normalized currents provided to the voice coils against a modulation coefficient a;

[0019] FIG. 7, in a graph, plots relative increase in the power dissipated as a function of the modulation coefficient a;

[0020] FIG. 8, in a functional schematic view, illustrates a power amplifier in accordance with a further embodiment of the invention;

[0021] FIG. 9, in a graph, plots the time each power transistor is on as a function of the modulation coefficient a;

[0022] FIG. 10, in a graph, plots the switching frequency between the power transistors as a function of the modulation coefficient a for two different values of bias currents expressed by their normalized values a_b;

[0023] FIG. 11, in a graph, plots the relationship between duty cycle and the modulation coefficient a;

[0024] FIG. 12, in a graph, illustrates the polarities and relationships between the power transistor threshold currents and the modulation coefficient a;

[0025] FIG. 13, in a graph, plots power transistor drain voltages as a function of time;

[0026] FIG. 14, in a graph, plots a ratio of maximum drain voltage to supply voltage as a function of the modulation coefficient a;

[0027] FIG. 15, in a graph, plots examples of waveforms of speaker voice coil currents assuming a sinusoidal input signal;

[0028] FIG. 16, in a graph, illustrates the phenomena of cross-over distortions;

[0029] FIG. 17, in a graph, illustrates a method of compensating for cross-over distortions;

[0030] FIG. 18, in a functional schematic view, illustrates a power amplifier employing a circuit for cross-over distortion compensation in accordance with a further embodiment of the invention;

[0031] FIG. 19, in a detailed schematic view, illustrates the power amplifier of FIG. 18;

[0032] FIG. 20, in a graph, plots the power transistor drain voltages and shows voltage spikes caused by parasitic inductances of a switching transformer;

[0033] FIG. 21, in a graph, illustrates an example of control waveforms in the power amplifier of FIGS. 22 and 25;

[0034] FIG. 22, in a functional schematic view, illustrates a power amplifier class "A" in accordance with an embodiment of the invention;

[0035] FIG. 23, in a graph, illustrates the time on for each power transistor in the power amplifier of FIGS. 22 and 25 as a function of the modulation coefficient a;

[0036] FIG. 24, in a graph, plots an example of switching frequency in the power amplifier of FIGS. 22 and 25 as a function of the modulation coefficient a; and,

[0037] FIG. 25, in a detailed schematic view, illustrates the power amplifier of class "A" of FIG. 22.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS OF THE INVENTION

[0038] Referring to FIG. 1, there is illustrated in a schematic view, a speaker 100 with a dual split voice coil 102 in accordance with the prior art. Voltages U_1 and U_2 are applied to the extreme terminals 104, 106 of voice coil 102, while voltage E is applied to common terminal 103. The resulting currents i_{a1} and i_{a2} flow through the voice coil 102, which sits in a magnetic flux created by a magnet (not shown) within speaker 100. The current flowing through the voice coil 102 causes a force to act on the voice coil 102 according to the equation F=Bli where F is the force, B is the magnetic flux in which the wire is placed, l is the length of the wire and i is the current running through the wire (this equation assumes that the wire is perpendicular to the magnetic flux). The means by which the force F can be used to create sound are well known in the art.

[0039] The split voice coil 102 is in fact composed of two coils 108, 110. These coils may be split, dual, overlay or bifilar voice coils. The forces created by coils 108 and 110 respectively are given as:

$$F_1 = B_1 \cdot l_1 \cdot i_{a1} \qquad (1)$$

$$F_2 = B_2 \cdot l_2 \cdot (-i_{a2}) \qquad (2)$$

[0040] where B_1 and B_2 are the densities of magnetic flux within which each of the coils 108 and 110 is placed respectively, l_1 and l_2 are the lengths of the wires or conductors of each of the coils, and i_{a1} and i_{a2} are the currents through each of the coils. The second current is indicated as negative given that, as shown in FIG. 1, it is opposite in direction to the first current.

[0041] Typically, magnetic flux is not uniformly distributed along the voice coil. One way of dealing with this is to have B_1 and B_2 represent the average densities of magnetic flux along the length of the coil. Another way of dealing with this is to assume non-uniform values of magnetic flux density B_1 and B_2 and define each of l_1 and l_2 to be the effective length of its respective coil.

[0042] Symmetrical coils can be provided by two identical voice coils being positioned side-by-side, one coil being wound on top of another coil (overlay) or both coils made simultaneously using bifilar coils. Bifilar coils assure almost identical coils, but methods in which one coil is overlaid on top of another give very similar results.

[0043] For the case of two coils that are symmetrically positioned with respect to a magnetic field created by the magnet in the voice coil gap, the following relationships hold:

$$B_1 = B_2 = B \tag{3}$$

$$l_1 = l_2 = \frac{l}{2} \tag{4}$$

[0044] Thus, the total force resulting from both coils is

$$F = F_1 + F_2 = B \cdot l \cdot \frac{i_{a1} - i_{a2}}{2} \tag{5}$$

[0045] The voice coils must deliver the force required to move the speaker components. This force can be generated by one coil at a time, or by both simultaneously. That is, the force generated by the coils can result from an equal distribution of current between the two coils, by an uneven distribution of current between the two coils, or, in the extreme, by current being provided to only one coil at a time. The two extreme cases are considered below. In the first case, current is only provided to the first coil, such that i_{a2}=0. In the second case, the coils are equally active as equal amounts of current are supplied to each: i_{a1}=−i_{a2}=i_c, where i_c is a hypothetical current through both currents simultaneously that provides the same force as the sum of the forces generated by the actual component currents i_{a1} and i_{a2}. In both cases, the force F must be the same to get the same sound from the speaker. Thus, in the first case i_{a1}=2·i_c.

[0046] Current flowing through the coils' resistances dissipates heat. Each coil has a resistance equal to R/2. Thus, the power P_1 dissipated in the first coil can be expressed as follows:

$$P_1 = \frac{R}{2} \cdot (i_{a1})^2 + 0 = \frac{R}{2} \cdot (2 \cdot i_c)^2 = 2 \cdot R \cdot i_c^2 \tag{6}$$

[0047] Where equal currents are provided to each coil, the total power P_c dissipated can be calculated as follows:

$$P_c = \frac{R}{2} \cdot (i_c)^2 + \frac{R}{2} \cdot (i_c)^2 = R \cdot i_c^2 \tag{7}$$

[0048] Combining these two equation yields

$$P_1 = 2 P_c \tag{8}$$

[0049] From this equation, it is apparent that when a current is provided to only one coil, as opposed to being divided equally between both coils, the power dissipated is doubled. However, to complete this analysis, it is necessary to consider cases in which $i_{a1} \neq 0$ and $i_{a2} \neq 0$ Specifically, in these cases,

$$P_1 + P_2 = \frac{R}{2} \cdot (i_{a1}^2 + i_{a2}^2) \tag{9}$$

[0050] As current i_{a2} flows in the opposite direction to current i_{a1},

$$i_c = \frac{i_{a1} - i_{a2}}{2} \tag{10}$$

[0051] current i_c dissipates power P_c

$$P_c = R \cdot i_c^2 \tag{11}$$

[0052] Combining equations (9) and (11) gives

$$\frac{P_1 + P_2}{P_c} = \frac{i_{a1}^2 + i_{a2}^2}{2 \cdot i_c^2} \tag{12}$$

[0053] Combining equations (10) and (12) yields

$$\frac{P_1 + P_2}{P_c} = 2 - 2 \cdot \left(\frac{i_{a1}}{i_c}\right) + \left(\frac{i_{a1}}{i_c}\right)^2 \tag{13}$$

[0054] FIG. 2, in a graph, illustrates a normalized plot 112 of the ratio of the total power dissipated in both coils to the power dissipated by current i_c as a function of the ratio of the current through coil 110 to the current flowing through both coils, according to equation (13). The plot 112 indicates that the point of a minimum power dissipation is, as expected, at

$$\frac{i_{a1}}{i_c} = 1.$$

[0055] For example,

$$\text{if } \frac{i_{a1}}{i_c} = 0.5 \text{ or if } \frac{i_{a1}}{i_c} = 1.5,$$

[0056] in which case i_{a1}/i_{a2}=⅓ or i_{a1}/i_{a2}=3 respectively, then one will need 25% more power for the same force. Thus, in general, the more uniform the current distribution between the coils is, the higher the efficiency. High efficiency is particularly important in high power systems, where even a small decrease in efficiency translates to a large amount of dissipated heat.

[0057] U.S. Pat. No. 2,959,640 (Schultz) and U.S. Pat. No. 4,130,725 (Nagel) disclose the use of class B amplifiers for split coil transducers. Class B, push pull or other amplifiers alternate the current between coils at audio frequencies by delivering the current to half of the coils at a time with the other half being left inactive. According to the foregoing

analysis, this leads to a 50% reduction in the maximum possible efficiency, and to the above-described problem of increased heat dissipation.

[0058] Referring to **FIG. 3a,** there is illustrated in a schematic view, a full bridge class D amplifier **114** according to the prior art. The amplifier **114** includes a modulator **116** to which an input signal is applied. Based on the input signal provided, the modulator **116** provides signals to the field effect transistor (FET) driver **118,** which, in turn, drives the transistors **120, 122, 124, 126.** The transistors **120, 122, 124, 126** are powered by the power supply **128** and drive the speaker **130** through a filter comprising inductors **132** and **134,** as well as a capacitor **136.**

[0059] Referring to **FIG. 3b,** there is illustrated in a schematic view a half bridge class D amplifier **138** in accordance with the prior art. The amplifier **138** includes a modulator **140** to which an input signal is provided. Based on the input signal provided, the modulator **140** provides appropriate signals to the FET driver **142,** which, in turn, drives the transistors **144, 146.** However, unlike the full bridge configuration of **FIG. 3,** only two transistors **144, 146** are required. The filter of the amplifier **138** includes an inductor **148** and a capacitor **150.** Unlike the amplifier **114** of **FIG. 3a,** amplifier **138** requires a dual power supply **151.**

[0060] The amplifiers **114** and **138** of **FIGS. 3a** and 3b respectively suffer from the disadvantage of requiring high side FET drivers. These circuits are complicated and difficult to construct, especially for high power applications with highly reactive loads.

[0061] Referring to **FIG. 4,** there is illustrated in a schematic view, an amplifier **152** in accordance with an aspect of the present invention. As shown, the amplifier **152** includes a power supply voltage E, a branch **152a** and a branch **152b.** The branch **152a** includes a transistor **154,** having a source **154s,** gate **154g,** and drain **154d** and a half winding **156a** of a transformer winding **156** (which may be a high frequency transformer or an impulse transformer). Similarly, the branch **152b** includes a transistor **158b** having a source **158s,** gate **158g** and drain **158d,** and a half winding **156b** of the transformer **156.** Branch **152a** is connected to terminal **160a** of speaker coil **162a,** while branch **152b** is connected to speaker terminal **160b** of speaker coil **162b.** The terminals of the transformer **156** are connected such that the two currents provided by the transistors **154** and **158** flowing in the same direction in each branch would enter the transformer through terminals **157a** and **157b** respectively, which are of opposite polarity.

[0062] The transformer **156** has a 1:1 turn ratio and an inductance L. As shown, it is connected in series between the transistor drains **154d** and **158d** and the terminals **160a** and **160b** of the split voice coil speaker **164.** The transistor drains **154d** and **158d** are connected to transformer terminals of different polarities. Transistors **154** and **158** are switched on and off alternatively upon reaching the threshold currents i_1 and i_2 respectively. These currents are sensed using resistors **166a** and **166b** of branches **152a** and **152b** respectively, which are connected in series with their associated transistor source terminals **154s** and **158s** respectively.

[0063] Amplifier **152** also includes filtering capacitors **168a** and **168b,** which are sufficiently large such that voltages U_1 and U_2 at the speaker terminals at **160a** and **160b** remain relatively constant during switching intervals T_1 and T_2.

[0064] The power transistors **154** and **158** alternately allow current to flow through each of the branches according to the input signal. This, in turn, energizes each of the coils **162a** and **162b** to thereby produce sound. The transformer **156** stores the energy produced by the conducting branch and transfers it to the other branch when the transistor of the conducting branch switches off and that of the non-conducting branch switches on. When a voltage source is connected across the terminals of an inductor, the inductor's current increases linearly according to the following equation:

$$U = L \cdot \frac{di}{dT} \tag{14}$$

[0065] where U is the applied voltage, di is the increment of current, and dT is the increment of time.

[0066] Referring to **FIG. 5,** the waveforms of current through each transistor **154** and **158** are illustrated. When the transistor **154** is turned off, current from the inductor **156** starts to flow through transistor **158.** The absolute value of the current within the inductor does not change during this transition. Instead, only its direction changes due to the way the transformer **156** is connected to the circuit. The same happens during the next transition when transistor **158** turns off and transistor **154** turns on. Thus, as shown in **FIG. 5,** the initial current for transistor **154** is $-i_2$ and the initial current for transistor **158** is $-i_1$. Referring back to equation (14), the following equations for the voltages U_1 and U_2 at speaker terminals **160a** and **160b** respectively can be determined:

$$U_1 = L \cdot \frac{i_1 - (-i_2)}{T_1} = L \cdot \frac{i_1 + i_2}{T_1} \tag{15}$$

$$U_2 = L \cdot \frac{i_2 - (-i_1)}{T_2} = L \cdot \frac{i_2 + i_1}{T_2} \tag{16}$$

[0067] From these equations, we derive that

$$U_1 \cdot T_1 = U_2 \cdot T_2 \tag{17}$$

[0068] The drain currents of the switching transistors **154** and **158** have high frequency components and DC components. The high frequency components are buffered by capacitors **168a** and **168b** such that only the DC or average values i_{a1} and i_{a2} of the drain currents reach the speaker coils **168a** and **168b.** These average values are given by the following equations:

$$i_{a1} = \frac{i_1 - i_2}{2} \cdot \frac{T_1}{T_1 + T_2} \tag{18}$$

$$i_{a2} = \frac{i_1 - i_2}{2} \cdot \frac{T_2}{T_1 + T_2} \tag{19}$$

[0069] $T_1 + T_2$ is the period of the currents i_1 and i_2. These equations were derived by assuming triangular waveforms of currents as shown in **FIG. 5.** Specifically, referring to **FIG. 5,** waveform **170** is the current through the source of power transistor **154,** while waveform **172** is the current through the source of power transistor **158.** Waveforms **170**

5

and **172** would typically be at frequencies between 100 and 200 kilohertz. Lower frequencies could be used, but this will require larger inductors. Further, the frequency should not be so low that these waveforms are audible. Higher frequencies may also be used, depending on technologies available. Similarly, stippled line **174** designates the average current value i_{a1} actually supplied to the coil **168***a*, and stippled line **176** designates the average current i_{a2} actually supplied to the speaker coil **168***b*. Typically, waveforms **174** and **176** would have frequencies in the audible range of 20 hertz to 20 kilohertz, although for some speaker systems, waveforms **174** and **176** may have frequencies below 20 hertz—the pressure waves generated by these waveforms would not be heard, but could be felt.

[0070] As shown in **FIG. 5**, during period T_1 transistor **154** conducts current while transistor **158** does not. As shown in **FIG. 5**, this situation is reversed during time period T_2 during which transistor **158** conducts while transistor **154** does not. As shown in **FIG. 5**, the frequency of current waveforms **170** and **172** is much higher than the frequency of current waveforms **174** and **176**. Due to this fact, the currents i_{a1} and i_{a2} which represent audio signals appear to be constant in **FIG. 5**; however, if the X axis of **FIG. 5** were extended over a suitably long period of time, these audio signals would be seen to vary in accordance with the audio content of the signal. These audio signals are also constant and concurrently provided over the period of waveforms **170** and **172**, despite the fluctuations in current waveforms **170** and **172**, due to the capacitors **168***a* and **168***b* that filter the high frequency component of the signal out before it reaches the coils **162***a* and **162***b*. The currents i_{a1} and i_{a2} are also oriented in the coils **162***a* and **162***b* respectively such that both currents i_{a1} and i_{a2} constructively contribute to the force provided to the voice coil **162**. That is, given the configuration of the coils **162***a* and **162***b* in **FIG. 4**, the currents i_{a1} and i_{a2} have opposite polarities as shown in **FIG. 5**. This is also shown by equation (18) and (19).

[0071] In the discussion that follows, i_0 is defined to be the difference between i_1 and i_2. That is

$$i_1 - i_2 = i_i \qquad (20)$$

[0072] Substituting this equation into equations (18) and (19), yields

$$i_{a1} = \frac{i_0}{2} \cdot \frac{T_1}{T_1 + T_2} \qquad (21)$$

$$i_{a2} = -\frac{i_0}{2} \cdot \frac{T_2}{T_1 + T_2} \qquad (22)$$

[0073] Combining equations (21) and (22) yields the relation

$$i_{a1} - i_{a2} = \frac{i_0}{2} \qquad (23)$$

[0074] The flows of these average currents through voice coils **168***a* and **168***b*, each of which has resistance R/2, create the following voltage drops respectively:

$$E - U_1 = \frac{R}{2} \cdot i_{a1} \qquad (24)$$

$$E - U_2 = \frac{R}{2} \cdot i_{a2} \qquad (25)$$

[0075] where E is the power supply voltage. Solving for U_1 and U_2 using the above equations (21), (22), (23), (24) and (25), yields

$$U_1 = E - \frac{R \cdot i_0}{4} \cdot \frac{T_1}{T_1 + T_2} \qquad (26)$$

$$U_2 = E + \frac{R \cdot i_0}{4} \cdot \frac{T_2}{T_1 + T_2} \qquad (27)$$

[0076] The above, equations (17), (26) and (27) can be solved. This solution can be simplified by introducing a modulation coefficient a, where

$$\frac{R \cdot i_0}{4 \cdot E} = a \qquad (28)$$

[0077] Substituting the modulation coefficient a into equations yields

$$U_1 = \frac{E}{2} \cdot \left(1 - a + \sqrt{1 - a^2}\right) \qquad (29)$$

$$U_2 = \frac{E}{2} \cdot \left(1 + a + \sqrt{1 - a^2}\right) \qquad (30)$$

[0078] Clearly, a must conform to the following condition

$$-1 \leq a \leq 1 \qquad (31)$$

[0079] The modulation coefficient a expresses the degree of achieving maximum possible values during amplifier operations. From equations (24), (25), (29) and (30), the following relations can be derived

$$i_{a1} = \frac{E}{R} \cdot \left(1 + a - \sqrt{1 - a^2}\right) \qquad (32)$$

$$i_{a2} = \frac{E}{R} \cdot \left(1 - a - \sqrt{1 - a^2}\right) \qquad (33)$$

[0080] The amplifier of **FIG. 4** will work with a wide range of loads and supply voltages. Accordingly, the foregoing equations can be normalized as follows:

$$i_{a1N} = 1 + a - \sqrt{1 - a^2} \qquad (34)$$

$$i_{a2N} = 1 - a - \sqrt{1 - a^2} \qquad (35)$$

[0081] Referring to **FIG. 6**, these normalized currents are plotted in a graph in which normalized current is plotted against the modulation coefficient a. That is, lines **178** and **180** indicate the currents through coils **162***a* and **162***b*

respectively. Line **182** designates the current that would flow through both coils in order to generate the same force.

[0082] Recall from equations (3) and (10) above that

$$F = F_1 + F_2 = B \cdot l \cdot i_c \qquad (36)$$

[0083] From equations (10), (32) and (33) it follows that

$$i_c = \frac{E}{R} \cdot a \qquad (37)$$

[0084] The term E/R is common to equations (32), (33) and (37). From the foregoing, it can be deduced that for the normalized value i_{cN} of current i_c,

$$i_{cN} = a \qquad (38)$$

[0085] This relation is represented by straight line **182** in **FIG. 6**.

[0086] Current i_c in its normalized version i_{cN} is a hypothetical current through both coils simultaneously that provide the same force as the sum of the forces generated by the actual component currents i_{a1} and i_{a2}.

[0087] From equations (37), (36), (28) and (20), the following equation can be derived:

$$F = B \cdot l \cdot \frac{i_1 - i_2}{4} \qquad (39)$$

[0088] This equation shows that the force F generated by the voice coil assembly is a linear function of a or i_1-i_2. That is, despite the fact that the individual voice coil currents are non-linear functions of the input signal as shown in **FIG. 6**, their sum is perfectly linear as shown by equation (39). If the difference between the threshold currents i_1 and i_2 is controlled to correspond to the input audio signal, the system response will be clean and without distortion.

[0089] Ideally, switching power amplifiers do not experience any energy losses. This is also the case of the amplifier according to the present invention. If that is the case, the total power P_t dissipated in both coils can be represented by the equation

$$P_t = P_1 + P_2 = E \cdot (i_{a1} + i_{a2}) \qquad (40)$$

[0090] Combining this equation with equations (32) and (33) yields

$$P_t = \frac{E^2}{R} \cdot 2 \cdot (1 - \sqrt{1 - a^2}) \qquad (41)$$

[0091] The power P_c dissipated by the current i_c is

$$P_c = \frac{E^2}{R} \cdot a^2 \qquad (42)$$

[0092] The ratio

$$\frac{P_t}{P_c}$$

[0093] is then given by the following equation:

$$\frac{P_t}{P_c} = 2 \cdot \frac{1 - \sqrt{1 - a^2}}{a^2} \qquad (43)$$

[0094] A similar result could be derived from equations (13), (32) and (37) or (13), (34) and (38).

[0095] Recall that equation (13) is

$$\frac{P_1 + P_2}{P_c} = 2 - 2 \cdot \left(\frac{i_{a1}}{i_c}\right) + \left(\frac{i_{a1}}{i_c}\right)^2$$

[0096] This equation can be used for any signal provided the substitute currents are expressed by their root mean square (RMS) values and all have the same shape. It can also be used in the case of the present amplifier provided that all the currents are DC currents. Accordingly, equation (43) is valid for a DC signal only.

[0097] Referring to **FIG. 7**, the ratio

$$\frac{P_t}{P_c}$$

[0098] is plotted against the modulation coefficient a according to equation (43). This is line **184** in the graph. The power dissipation is shown relative to the power dissipated when both coils are conducting evenly. Thus, as can be seen from curve **186** of graph **188**, which represents power dissipation by class B amplifiers, class B amplifiers always dissipate twice the power due to the fact that only one coil is conducting at a time. Curve **184**, determined by equation (43), indicates the power dissipated when the input signal is a DC signal. Curve **190** indicates the power dissipated when the signal is sinusoidal.

[0099] Equation (43) is not applicable in the case of a sinusoidal signal. Accordingly, the appropriate powers must be calculated using equations (41) and (42). Modulation coefficient a has to be substituted by a(t)=a·sin(t). Substituting this equation yields

$$P_t = \frac{E^2}{R} \cdot \frac{2}{\pi} \cdot \int_0^\pi \left(1 - \sqrt{1 - (a \cdot \sin(t))^2}\right) dt \qquad (44)$$

$$P_c = \frac{E^2}{R} \cdot \frac{1}{\pi} \cdot \int_0^\pi (a \cdot \sin(t))^2 dt = \frac{E^2}{R} \cdot \frac{a^2}{2} \qquad (45)$$

[0100] Combining these equations yields

$$\frac{P_t}{P_c} = 2 \cdot \frac{\int_0^\pi \left(1 - \sqrt{1-(a\cdot \sin(t))^2}\,\right)dt}{\int_0^\pi (a\cdot \sin(t))^2 dt} = \tag{46}$$

$$\frac{4}{\pi \cdot a^2} \cdot \int_0^\pi \left(1 - \sqrt{1-(a\cdot \sin(t))^2}\,\right)dt$$

[0101] Plotting the curve yielded by equation (46) provides line **190** of **FIG. 7**.

[0102] The advantages of the amplifier of **FIG. 4** are readily apparent from **FIG. 7**. If the modulation for the sinusoidal signal reaches 0.8, then the additional power dissipation is only 17% of the power dissipation that would result if both coils were working uniformly—i.e. if equal currents were being supplied to both coils. By contrast, with class B and other amplifiers in which coils are powered alternately, the additional power dissipation will always be 100% of what the powered dissipation would be if both coils were working uniformly, regardless of the value of the modulation coefficient. However, **FIG. 7** compares the losses of power in the speaker only. In addition to these losses, class B amplifiers will have their own losses and limited efficiency, unlike switching amplifiers, which enjoy efficiencies close to 100%.

[0103] As is apparent from **FIG. 5**, the amplifier **152** operates by switching the power transistors alternatively ON and OFF whenever their currents reach certain values. As a result, the amplifier **152** is self-oscillating and the switching frequency is related to the values of the amplifier components.

[0104] Recall that modulation coefficient a determined by the following equations:

$$a = \frac{R\cdot i_0}{4\cdot E} = \frac{R\cdot (i_1 - i_2)}{4\cdot E} \tag{47}$$

[0105] The value of the modulation coefficient a can be positive or negative and varies according to the input signal. However, the threshold currents i_1 and i_2 must always be positive. Accordingly, additional circuitry must be added to the amplifier **152** for this amplifier to operate properly. For the purpose of this analysis, it is assumed that while one of the currents i_1 or i_2 varies, the other remains constant as shown in **FIG. 12**. The initial or biasing current i_b and its normalized value a_b are then related by equation (48) similar to equation (28)

$$a_b = \frac{R\cdot i_b}{4\cdot E} \tag{48}$$

[0106] Algebraic transformation of equations (15), (16), (20), (29), (30), (28) and (48), and substitution of one of the currents by current i_b lead to equations (49) and (50) for switching times T_1 and T_2:

$$T_1 = \frac{8\cdot L}{R}\cdot \frac{a+2\cdot a_b}{1-a+\sqrt{1-a^2}} \tag{49}$$

$$T_2 = \frac{8\cdot L}{R}\cdot \frac{a+2\cdot a_b}{1+a+\sqrt{1-a^2}} \tag{50}$$

[0107] From equations (49) and (50), the switching frequency f_s and duty cycle D can be determined:

$$f_s = \frac{1}{T_1+T_2} = \frac{R\cdot \sqrt{1-a^2}}{8\cdot L\cdot (|a|+2\cdot a_b)} \tag{51}$$

$$D = \frac{T_1}{T_1+T_2} = \frac{\sqrt{1+a}}{\sqrt{1+a+\sqrt{1-a}}} \tag{52}$$

[0108] Referring to **FIG. 8**, there is illustrated in a functional schematic view, an amplifier **200** in accordance with a preferred embodiment of the invention. As illustrated, the amplifier **200** consists of two main branches **200a** and **200b**, each of which processes either the positive or negative portion of the input signal. An input signal V_1 is fed to the inputs of two very similar circuits, called the Positive Signal Circuit (PSC) **202a** and Negative Signal Circuit (NSC) **202b**. The output voltage of the PSC **202a** equals its input voltage if the input voltage is positive and equals zero if the input voltage is negative. The output voltage of the NSC **202b** equals zero if the input voltage is positive and equals the inverted input voltage when the input voltage is negative. The output of PSC **202a** is connected to summing circuit **204a**, while the output of NSC **202b** is connected to summing circuit **204b**. The summing circuits **204a** and **204b** add constant and positive voltage V_b to the output voltages of the PSC **202a** and the NSC **202b**, thereby generating voltages V_{i1} and V_{i2} respectively. The relationship between voltages V_{i1} and V_{i2}, and input voltage V_i, is described by equations (54) and (55):

$$V_{i1} = \frac{1}{2}(|V_i| + Vi) + V_b \tag{53}$$

$$V_{i2} = \frac{1}{2}(|V_i| - Vi) + V_b \tag{54}$$

[0109] Voltages V_{i1} and V_{i2} are always positive and their minimal values equals V_b. A graph plotting equations (53) and (54) would look quite similar to the graph of **FIG. 12**.

[0110] Referring back to **FIG. 8**, each branch of the amplifier circuit **200** includes a voltage comparator **206a** and **206b**. Comparator **206a** compares voltages V_{i1} and the voltage across a sensing resistor **208a**, while comparator **206b** compares voltages V_{i2} and the voltage across sensing resistor **208b**. The operation of voltage comparators are known by those of skill in the art. The current sensing resistors **208a** and **208b** convert transistor currents to voltages. If the drain current of transistor **210a** is higher than voltage V_{i1} divided by the value of resistor **208a**, then the output of comparator **206a** is the binary output one. Con-

trariwise, if the current obtained by dividing the voltage V_{i1} by the value of the resistance of resistor $208a$ is higher than the drain current of transistor $210a$, then the comparator outputs zero. Similarly, comparator $206b$ outputs one when the drain current of transistor $210b$ is higher than voltage V_{i2} divided by the resistance of resistor $208b$, and outputs a zero when the drain current of transistor $210b$ is lower than voltage V_{i2} divided by the resistance of resistor $208b$.

[0111] Each branch $200a$ and $200b$ of the amplifier 200 also includes a NOR gate $212a$ and $212b$ respectively. The NOR gates $212a$ and $212b$ are cross-coupled so as to produce an R-S flip-flop 214. The R-S flip-flop has two complementary outputs. Say that the output connected to the gate of transistor $210a$ represents logical one and the output to transistor $210b$ represents logical zero. The drain current of transistor $210a$ then increases linearly and at some point of time reaches value $i_1 = V_{i1}/R_1$. This occurrence is illustrated in the vertical shifts of **FIG. 5**. At this point of time, the output signal of comparator $206a$ changes from zero to one. This changes the state of the R-S flip-flop 214 and turns the transistor $210a$ OFF and transistor $210b$ ON. The state of R-S flip-flop 214 remains constant after this until the rising current of transistor $210b$ crosses the value i_2 determined by V_{i2} and the resistance of resistor $208b$, and then the cycle repeats—i.e. the output of comparator $206b$ changes from zero to one, thereby changing the state of the R-S flip-flop and turning the transistor $210a$ ON and the transistor $210b$ OFF. The entire circuit 200 oscillates with the frequency given by equation (51) above, and peak currents i_1 and i_2 vary with the input signal (**FIG. 12**).

[0112] The current sensing resistors $208a$ and $208b$ should have identical resistances. If this is the case, then

$$i_1 = \frac{V_{i1}}{R_1} \tag{55}$$

$$i_2 = \frac{V_{i2}}{R_1} \tag{56}$$

[0113] From equations (39) and (53) to (56), the following equality can be derived:

$$F = B \cdot l \cdot \frac{V_i}{4 \cdot R_1} \tag{57}$$

[0114] This equation shows that the force generated by the speaker motor assembly is a linear function of the input voltage V_i. The current sensing circuitry controls the power transistors $210a$ and $210b$ so that the amplifier 200 works as a current amplifier as indicated by equation (57). **FIG. 9** illustrates in a graph the switching times for power transistors $210a$ and $210b$ as a function of the modulation coefficient a. The vertical axis shows time ON and is expressed as a multiple of $8L/R$, where L is inductance of the transformer 216 and R is the total resistance of the coil 218. Curve 230 represents the time ON for transistor $210a$, while curve 232 illustrates the time ON for power transistor $210b$.

[0115] **FIG. 10**, in a graph, illustrates the switching frequency as a function of the absolute value of modulation

coefficient a. The vertical axis is expressed as multiples of R/8L, similar to **FIG. 9**. Curve 234 represents the switching frequency for a biasing current having a normalized value of 0.05. Curve 236 represents the switching frequency for a biasing current having a normalized value of 0.1. From curves 234 and 236 of **FIG. 10** and from curves 230 and 232 of **FIG. 9**, it is apparent that both times T_1 and T_2 increase, and the switching frequencies decrease, when the signal value increases. The biasing current i_b has the highest influence on times and frequency when the input signal is small. The duty cycle is independent of i_b.

[0116] When one of the transistors is ON, the other one is OFF and its drain voltage equals VD_M. With no input signal, modulation coefficient a equals zero, and VD_M is twice as high as the power supply voltage E. The exact value of VD_M as function of the modulation coefficient a is given by the following equation:

$$VD_M = U_1 + U_2 = E \cdot (1 + \sqrt{1-a}) \tag{58}$$

[0117] The function described by this equation is plotted as curve 260 in **FIG. 14**. VD_M should not be higher than the maximum rated voltage of the chosen transistors.

[0118] The present analysis was conducted, and the equations were derived on the basis of the assumption that the speaker impedance does not vary with frequency. In actuality, this is not the case. However, equation (39) is valid notwithstanding this. As a consequence, the force generated by the motor structure of the speaker depends only on the difference between currents i_1 and i_2, and, as shown by equation (57), depends on input voltage V_i alone. An amplifier in accordance with an embodiment of the present invention consequently behaves like a normal current mode power amplifier in responding to variations in speaker impedance.

[0119] Equations (53), (54), (55) and (56) define the relationship between threshold currents i_1 and i_2, and input signal V_1 represented by modulation coefficient a. This relationship is also illustrated by the graph of **FIG. 12**. Specifically, **FIG. 12** illustrates the function of PSC $202a$, NSC $202b$, summing circuit $204a$, and summing circuit $204b$ of **FIG. 8**, and a bias voltage V_b. Bias voltage V_b is responsible for generating the biasing current i_b. From **FIG. 12**, it is apparent that current i_0 varies linearly with modulation a and input signal.

[0120] Referring to **FIG. 13**, there is illustrated in a graph the relationship between transistor drain voltages VD_1 and VD_2, voltages at the speaker terminals U_1 and U_2, maximum drain voltage VD_m and supply voltage E as a function of time. From **FIG. 13** it is apparent that while voltage U_1 is smaller than supply voltage E, voltage U_2 is larger than supply voltage E; however, all voltages are smaller than 2 E.

[0121] Referring to **FIG. 15**, there is illustrated in a graph examples of waveforms of speaker currents i_{a1} and i_{a2} through each coil, assuming sinusoidal audio input signals and modulation coefficient 0.9. **FIG. 15** shows that despite individual coil currents being distorted, the difference between the individual coil currents, which is responsible for generating the force, is sinusoidal and free from distortion. This graph is analogous to the graph of **FIG. 6**; the one being for a instantaneous values of an audio input signal while the other is for a sinusoidal audio input signal as a function of time.

[0122] Referring back to **FIG. 8**, the operation of the PSC **202***a* and the NSC **202***b* resemble the operation of a halfway rectifier. That is, the output signal represents only the positive or negative half cycles of an applied sinusoidal input signal. The amplifier **200** uses two such circuits. It is particularly important that the characteristics of the PSC **202***a* and NSC **202***b* be closely matched and appropriately shaped. If not, then the amplified signal will have distortions. Consider the example of a simple diode and resistor circuit. The diode needs a certain bias voltage to conduct current. This means that only voltages higher than the bias voltage will appear at the output. This is a typical mechanism for generating so called crossover distortions as illustrated in **FIG. 16**.

[0123] Referring to **FIG. 16**, there is illustrated a sinusoidal signal represented by curve **300**, which is the input signal applied to the amplifier. Curves **302** and **304** are the process input signals applied to the inputs of comparators **206***a* and **206***b* respectively. The resulting signal is represented by curve **306**. Crossover distortions occur because the diodes used to rectify the waveforms require a minimum voltage to begin conducting. One way around this problem is to use NSC **202***b* to input into one of the comparators **206***a* or **206***b*, while inputting the sum of the original input signal and the output signal of NSC **202***b* into the other comparator **206***b* or **206***a* (**FIG. 18**).

[0124] Referring to **FIG. 17**, there are illustrated voltage forms, which have been compensated for crossover distortions. In **FIG. 17**, curve **308** is the input signal applied to the amplifier, and curves **310** and **312** are the process input signals applied to the inputs of comparators **206***a* and **206***b* respectively. The resulting signal applied to the amplifier circuit is shown as curve **314**. As can be seen, the crossover distortions have been eliminated.

[0125] The value of the bias voltage V_b, and, as a result, the value of the bias current i_b has no affect on the amplifier output signal. If the bias current i_b is very small, and there is no input signal, then the amplifier switching occurs at a very high frequency. On the other hand, if i_b is large, then the amplifier oscillates at low frequency, but the switching current is large. Thus, the optimal bias current will be a compromise between these two extremes. According to an embodiment of the invention, the bias current is selected to be around 10% of the modulation coefficient a_b determined by equation (48). However, this choice of bias current value is not critical.

[0126] Referring to **FIG. 18**, there is illustrated in a functional schematic view, an amplifier **200'** in accordance with a preferred embodiment of the invention. For clarity, the same reference numerals together with an apostrophe are used to designate elements analogous to those described above in connection with **FIG. 8**. As illustrated, the amplifier **200'** consists of two main branches **200***a''* and **200***b'*, each of which processes either the positive or negative portion of the input signal. An input signal V_i is fed to the inputs of two very similar circuits, called the Positive Signal Circuit (PSC) **322** and Negative Signal Circuit (NSC) **202***b'*. The output voltage of the NSC **202***b'* equals zero if the input voltage is positive and equals the inverted input voltage when the input voltage is negative. The PSC **322** consists of a summing junction **322***a* the inputs of which are the input signal V_i as well as the output of NSC **202***b'*. The input/

output characteristics of PSC **322** are similar to that of PSC **202***a* of **FIG. 8**. When the input signal is negative the output of NSC **202***b'* is positive but equal in magnitude. Thus, the output of PSC **322** is 0. When the input signal is positive the output of NSC **202***b'* is 0 and the output of PSC **322** is simply the input signal. The output of PSC **302** is connected to summing circuit **204***a'*, while the output of NSC **202***b'* is connected to summing circuit **204***b'*. The summing circuits **204***a'* and **204***b'* add constant and positive voltage V_b to the output voltages of the PSC **322** and the NSC **202***b'*, thereby generating voltages V_{i1} and V_{i2} respectively. The relationship between voltages V_{i1} and V_{i2}, and input voltage V_i, is described by equations (54) and (55) which were given above and are reproduced below:

$$V_{i1} = \frac{1}{2}(|V_i| + Vi) + V_b \tag{53}$$

$$V_{i2} = \frac{1}{2}(|V_i| - Vi) + V_b \tag{54}$$

[0127] As explained previously, voltages V_{i1} and V_{i2} are always positive and their minimal values equal V_b. A graph plotting equations (53) and (54) would look quite similar to the graph of **FIG. 12**.

[0128] Referring back to **FIG. 18**, each branch **200***a'* and **200***b'* of the amplifier circuit **200'** includes a voltage comparator **206***a'* and **206***b'* respectively. Comparator **206***a'* compares voltages V_{i1} and the voltage across a sensing resistor **208***a'*, while comparator **206***b'* compares voltages V_{i2} and the voltage across sensing resistor **208***b'*. The operation of voltage comparators are known by those of skill in the art. The current sensing resistors **208***a'* and **208***b'* convert transistor currents to voltages. If the drain current of transistor **210***a'* is higher than voltage V_{i1} divided by the value of resistor **208***a'*, then the output of comparator **206***a'* is the binary output one. Contrariwise, if the current obtained by dividing the voltage V_{i1} by the value of the resistance of resistor **208***a'* is higher than the drain current of transistor **210***a'*, then the comparator outputs zero. Similarly, comparator **206***b'* outputs one when the drain current of transistor **210***b'* is higher than voltage V_{i2} divided by the resistance of resistor **208***b'*, and outputs a zero when the drain current of transistor **210***b'* is lower than voltage V_{i2} divided by the resistance of resistor **208***b'*.

[0129] Each branch **200***a'* and **200***b'* of the amplifier **200'** also includes a NOR gate **212***a'* and **212***b'* respectively. The NOR gates **212***a'* and **212***b'* are cross-coupled so as to produce an R-S flip-flop **214'**. Each NOR gate has two complementary outputs. Say that the output connected to the gate of transistor **210***a'* represents logical one and the output to transistor **210***b'* represents logical zero. The drain current of transistor **210***a'* then increases linearly and at some point of time reaches value $i_1 = V_{i1}/R_1$. At this point of time, the output signal of comparator **206***a'* changes from zero to one. This changes the state of the R-S flip-flop **214'** and turns the transistor **210***a'* OFF and transistor **210***b'* ON. The state of R-S flip-flop **214'** remains constant after this until the rising current of transistor **210***b'* crosses the value i_2 determined by V_{i2} and the resistance of resistor **208***b'*, and then the cycle repeats—i.e. the output of comparator **206***b'* changes from zero to one, thereby changing the state of the R-S flip-flop

10

and turning the transistor **210***a*' ON and the transistor **210***b*' OFF. The entire circuit **200**' oscillates with the frequency given by equation (51) above, and peak currents i_1 and i_2 vary with the input signal.

[0130] The current sensing resistors **208***a*' and **208***b*' should have identical resistances. If this is the case, then

$$i_1 = \frac{V_{i1}}{R_1} \qquad (55)$$

$$i_2 = \frac{V_{i2}}{R_1} \qquad (56)$$

[0131] From equations (39) and (53) to (56), the following equality can be derived:

$$F = B \cdot l \cdot \frac{V_i}{4 \cdot R_1} \qquad (57)$$

[0132] This equation shows that the force generated by the speaker motor assembly is a linear function of the input voltage V_i. The current sensing circuitry controls the power transistors **210***a*' and **210***b*' so that the amplifier **200**' works as a current amplifier as indicated by equation (57).

[0133] Referring to **FIG. 19**, there is illustrated in a detailed schematic diagram an amplifier **400** corresponding to the amplifier **200**' of **FIG. 18**. Transistors **402***a* and **402***b* are power-switching transistors, which are connected to transformer **404**. That is, the drain of each transistor **402***a*, **402***b* is connected to a different winding of transformers **404**. In addition, the terminals of transformer **404** to which the drains or transistors **402***a*, **402***b* are attached, are of opposite polarity. The other terminals of transformer **404** are connected to the speaker coils, **406***a* and **406***b*, as well as to filtering capacitors **408***a* and **408***b*.

[0134] Resistors **410***a* and **410***b* are current sensing resistors connected in series with the source terminals of transistors **402***a* and **402***b*. These resistors provide a voltage at the source terminal that is proportional to the current passing through the sources of the transistors **402***a* and **402***b*. Small signal transistors **412***a* and **414***a*, in typical totem pole configuration, drive the gate of power transistors **402***a* through resistor **416***a*. Similarly, small signal transistors **412***b* and **414***b* drive the gate of power transistor **402***b* through resistor **416***b*. Power transistor **402***a* and **402***b* are turned OFF and ON in an alternating fashion when the current through the source of the conducting transistor reaches a threshold value. Thus, there are essentially two states of the circuit, one in which transistor **402***a* is conducting and the other in which transistor **402***b* is conducting. As there are periods of time between the switching of transistors in which it is necessary to maintain the current conditions, as well as the signals that affect the current state, a memory storage unit is required to store the current state of the circuit. An R-S flip-flop provides a memory unit for storing the current state of the circuit. The R-S flip-flop comprises voltage comparators **418***a*, **420***a*, **418***b* and **420***b*, which in practice may be all on the same integrated circuit chip.

[0135] Each of the comparators **420***a* and **420***b* compare a weighted sum of the current through the source of one of the power transistors **402***a* and **402***b* and the input voltage with a reference voltage. The reference voltages are adjusted by two pairs of voltage dividing resistors **422***a* and **424***a*, and resistors **422***b* and **424***b*. Preferably, the resistance of resistor **422***a* equals the resistance of resistor **422***b*, and the resistance of resistor **424***a* equals the resistance of resistor **424***b*, thereby causing the two reference voltages to be equal provided the resistors **424***a* and **424***b* are connected to the same voltage potential.

[0136] The resistors **426***a*, **428***a* and **430***a*, and resistors **426***b*, **428***b* and **430***b*, along with resistors **410***a* and **410***b*, are scaling resistors. The relative values of the resistances of these resistors establishes the relationship between the input signal, the threshold currents and the reference voltage. The appropriate resistances for each of the resistors can be determined from simple principles of voltage division. In practice, it is preferable to have the resistance of resistor **410***a* equal to the resistance of resistor **410***b*, the resistance of resistor **426***a* equal to the resistance of resistor **426***b*, and the resistance of resistor **430***a* equal to the resistance of resistor **430***b*, and the resistance of resistor **428***a* equal to the resistance of resistor **428***b*, thereby assuring that each comparator uses the same weighted sum of the input voltage, source current and reference voltage.

[0137] As described above, the structure of much of the amplifier circuit **400** is symmetrical. That is, one of the halves processes the negative portion of the input signal and the other processes the positive portion. To allow for a similar circuit design for both the negative and positive portions, it is desirable that the relevant portion of the input signals inputted into both branches be either only positive or only negative. This may be accomplished by inverting either the positive or negative portion of the input signal. In the amplifier **400**, the Positive Signal Circuit PSC **432** inverts the positive portion of the signal. Within the PSC, there is a buffer stage **434**, a half wave rectifier stage including diode **436** and resistor **438**, and a signal inverter including resistor **440** and operational amplifier **442**. Resistors **444***a*, **446***a*, **444***b* and **446***b* are pull up resistors that are necessary for the proper operation of the voltage comparators **418***a* and **418***b*. Capacitors **448***a* and **448***b* provide a feedback loop for the current sensing comparators between the output and non-inverting input. This allows for faster switching times and short dead time zones, which is necessary for the proper functioning of the R-S flip-flop. Capacitors **450***a*, **450***b*, **452***a* and **452***b* are used to reduce high frequency noise and to improve circuit stability.

[0138] Transformer **404** provides for energy storage and transfer. When one of the power transistors **402***a* or **402***b* is ON, energy is accumulated within the inductance of the transformer **404** as a result of current flowing through one of the windings of the transformer **404**. When the power transistor is turned OFF, and the other transistor is switched on, the energy accumulated in the inductance is released into the second winding as an electrical current.

[0139] It is important to achieve effective coupling between both transformer windings **404***a* and **404***b* in order to maximize efficiency and limit parasitic inductances that can lower the efficiency of energy transfer and cause voltage spikes in the drain of the switched OFF power transistor. If

the voltage spike reaches the transistor breakdown voltage, then this additional energy is discharged through the transistor, dissipating additional heat. This heat energy dissipated should not be higher than the rated avalanche energy of the transistor. However, in high power applications, it is often difficult to limit this energy to below the avalanche energy of the transistor. To compensate for this, the amplifier **400** includes a protective clamping circuit comprising clamping diodes **454a** and **454b,** capacitors **456a** and **456b,** resistors **458a** and **458b,** as well as a discharge circuit, including transistor **460,** resistor **462,** capacitor **464** and zener diodes **466, 468** and **470.** The clamping voltage should be less than the transistor breakdown voltage, and is selected by the zener diodes **466, 468** and **470.**

[0140] During normal operation, the drain voltages **490** and **492** of transistors **402a** and **402b** switch between about 0 volts and a voltage close to twice the power supply voltage to E. As a result, diodes **454a** and **454b** are alternately forward or reverse biased. Forward biased diode **454a** charges capacitor **456a** so the voltage, at the node where capacitor **466a,** diode **454a** and resistor **458a** are connected, is close to twice the supply voltage. This voltage remains relatively constant because capacitor **466a** stores energy during the time when diode **454a** is reverse biased. Similarly, forward biased diode **454b** charges capacitor **456b** so the voltage, at the node where capacitor **456b,** diode **454b** and resistor **458b** are connected, is close to twice the power supply voltage. This voltage remains relatively constant also due to the fact that the capacitor **456b** stores energy during the time when diode **454b** is reverse biased. Resistors **458a** and **458b** have very small values necessary to eliminate parasitic oscillations, which occur in any switching circuit. The other terminals of resistors **458a** and **458b** are connected together and to the collector of transistor **460.** Consequently, the value of the voltage at the collector of the transistor **460** remains relatively constant and close to twice the value of the power supply voltage.

[0141] Zener diodes **466, 468** and **470** are connected in series and between the collector and the base of transistor **460.** They can be replaced with a single diode having a break down voltage equal to the combined voltage of the diodes. However, in many cases lower voltage zener diodes are more readily available.

[0142] The combined value of the break down voltages of diodes **466, 468** and **470** is chosen to be slightly above the supply voltage E. On the other hand, the sum of combined values of the break down voltages of diodes **466, 468** and **470** and supply voltage E should be chosen to be below the break down voltage of transistors **402a** and **402b.** The emitter terminal of transistor **460** is connected to the source of supply voltage E. In consequence, the voltage across diodes **466, 468** and **470** remains below their break down voltages, and there is no current flow through resistor **462.** Base emitter voltage of transistor **460** is then 0, and the collector current of this transistor is also zero.

[0143] Sudden changes in the drain currents of transistors **402a** or **402b** caused by switching, together with parasitic inductances of transformer **404,** can generate voltage spikes **494,** which could damage those transistors. Voltage spikes **494,** as shown in **FIG. 20,** charge capacitors **456a** and **456b** to higher values and the voltage at the collector of transistor **460** rises. In consequence, the voltage across zener diodes

466, 468 and **470** reach their break down voltage, and current starts to flow through the resistor **462.** When the voltage across resistor **462** connected between the emitter and base of transistor **460** reaches about 0.7 volts, the transistor **460** starts to conduct a current. The further increase, even very small, of the voltage at the collector terminal of transistor **460** causes the collector current of this transistor to rise rapidly, never allowing this collector voltage and in consequence the drain voltages of transistors **454a** and **454b** to rise above twice the value of the supply voltage E. The current stored in the parasitic inductances of transformer **404** is then discharged through forward biased diodes **454a** or **454b,** and transistor **460** and transistors **402a** and **402b** are protected.

[0144] It is also possible to connect the emitter of transistor **460** to the circuit ground and to double the total break down voltage of the zener diodes. This solution works as well, but dissipates twice as much power in transistor **460** than is dissipated in the above-described example.

[0145] **FIG. 20** illustrates the voltages **490, 492** at the drains of transistors **402a** and **402b.** Also shown is the spike **494** reaching transistor break down voltage.

[0146] Referring to **FIG. 22,** there is illustrated in a functional schematic view, an amplifier **200"** in accordance with a preferred embodiment of the invention. For clarity, the same reference numerals together with two apostrophes are used to designate elements analogous to those described above in connection with **FIG. 8** and **18.** As illustrated, the amplifier **200"** consists of two main branches **200a"** and **200b",** each of which processes either the positive or negative portion of the input signal. An input signal V_i is applied to each branch **200a"** and **200b".** In branch **200a"** the signal is unaltered until reaching the summing junction **204a"** while in the branch **200b"** the signal is inverted by inverter **508** before reaching summing junction **204b".** The summing circuits **204a"** and **204b"** add constant and positive voltage V_b to the output voltage to the input signal and the output of the inverter **508,** thereby generating voltages V_{i1} and V_{i2} respectively. V_b is sufficiently large to ensure that the resulting voltages V_{i1} and V_{i2} are always positive. For a sinusoidal input signal, a graph plotting the voltages V_{i1} and V_{i2} would look quite similar to the graph of **FIG. 21.**

[0147] Referring back to **FIG. 22,** each branch of the amplifier circuit **200"** includes a voltage comparator **206a"** and **206b".** Comparator **206a"** compares voltages V_{i1} and the voltage across a sensing resistor **208a",** while comparator **206b"** compares voltages V_{i2} and the voltage across sensing resistor **208b".** The operation of voltage comparators are known by those of skill in the art. The current sensing resistors **208a"** and **208b"** convert transistor currents to voltages. If the drain current of transistor **210a"** is higher than voltage V_{i1} divided by the value of resistor **208a",** then the output of comparator **206a"** is the binary output one. Contrariwise, if the current obtained by dividing the voltage V_{i1} by the value of the resistance of resistor **208a"** is higher than the drain current of transistor **210a",** then the comparator outputs zero. Similarly, comparator **206b"** outputs one when the drain current of transistor **210b"** is higher than voltage V_{i2} divided by the resistance of resistor **208b",** and outputs a zero when the drain current of transistor **210b"** is lower than voltage V_{i2} divided by the resistance of resistor **208b".**

[0148] FIG. 21 illustrates in a graph the signals 498 and 500 within the amplifier 200". As shown the signals 498 and 500 are not rectified but are rather inverted with respect to each other with each having a DC signal added to it.

[0149] Each branch 200a" and 200b" of the amplifier 200" also includes a NOR gate 212a" and 212b" respectively. The NOR gates 212a" and 212b" are cross-coupled so as to produce an R-S flip-flop 214". Each NOR gate has two complementary outputs. Say that the output connected to the gate of transistor 210a" represents logical one and the output to transistor 210b" represents logical zero. The drain current of transistor 210a" then increases linearly and at some point of time reaches value $i_1 = V_{i1}/R_1$. At this point of time, the output signal of comparator 206a" changes from zero to one. This changes the state of the R-S flip-flop 214" and turns the transistor 210a" OFF and transistor 210b" ON. The state of R-S flip-flop 214" remains constant after this until the rising current of transistor 210b" crosses the value i_2 determined by V_{i2} and the resistance of resistor 208b", and then the cycle repeats—i.e. the output of comparator 206b" changes from zero to one, thereby changing the state of the R-S flip-flop and turning the transistor 210a" ON and the transistor 210b" OFF. The entire circuit 200" oscillates with the frequency given by equation (51) above, and peak currents i_1 and i_2 vary with the input signal. (FIG. 12).

[0150] The current sensing resistors 208a" and 208b" should have identical resistances. If this is the case, then

$$i_1 = \frac{V_{i1}}{R_1} \tag{55}$$

$$i_2 = \frac{V_{i2}}{R_1} \tag{56}$$

[0151] From equations (39) and (53) to (56), the following equality can be derived:

$$F = B \cdot l \cdot \frac{V_i}{4 \cdot R_1} \tag{57}$$

[0152] This equation shows that the force generated by the speaker motor assembly is a linear function of the input voltage V_i. The current sensing circuitry controls the power transistors 210a" and 210b" so that the amplifier 200" works as a current amplifier as indicated by equation (57).

[0153] FIG. 23 illustrates in a graph the switching times, for power transistor 210a" and 210b" of amplifier 200" of FIG. 22 as a function of modulation coefficient a. The vertical axis shows time ON and is expressed as a multiple of 8L/R, where L is inductance of the transformer 216" and R is the total resistance of the coil 218". Curve 504 represents the time ON for transistor 210a", while curve 506 illustrates the time ON for power transistor 210b". This figure is analogous FIG. 9 for amplifier 200.

[0154] FIG. 24 illustrates in a graph illustrates the switching frequency of amplifier 200" as a function of modulation coefficient a.

[0155] Referring to FIG. 25, there is illustrated in a detailed schematic diagram an amplifier 400' in accordance with a further preferred embodiment of the invention. For clarity, the same reference numerals together with an apostrophe are used to designate elements analogous to those described above in connection with FIG. 19. For brevity, some of the description of FIG. 19 is not repeated with respect to FIG. 25. Transistors 402a' and 402b' are power-switching transistors, which are connected to transformer 404'. That is, the drain of each transistor 402a', 402b' is connected to a different winding of transformers 404'. In addition, the terminals of transformer 404' to which the drains or transistors 402a', 402b' are attached, are of opposite polarity. The other terminals of transformer 404' are connected to the speaker coils, 406a' and 406b', as well as to filtering capacitors 408a' and 408b'.

[0156] Resistors 410a' and 410b' are current sensing resistors connected in series with the source terminals of transistors 402a' and 402b'. These resistors provide a voltage at the source terminal that is proportional to the current passing through the sources of the transistors 402a' and 402b'. Small signal transistors 412a' and 414a', in typical totem pole configuration, drive the gate of power transistors 402a' through resistor 416a'. Similarly, small signal transistors 412b' and 414b' drive the gate of power transistor 402b' through resistor 416b'. Power transistor 402a' and 402b' are turned OFF and ON in an alternating fashion when the current through the source of the conducting transistor reaches a threshold value. Thus, there are essentially two states of the circuit, one in which transistor 402a' is conducting and the other in which transistor 402b' is conducting. As there are periods of time between the switching of transistors in which it is necessary to maintain the current conditions, as well as the signals that affect the current state, a memory storage unit is required to store the current state of the circuit. An R-S flip-flop provides a memory unit for storing the current state of the circuit. The R-S flip-flop comprises voltage comparators 418a', 420a', 418b' and 420b', which in practice may be all on the same integrated circuit chip.

[0157] Each of the comparators 420a' and 420b' compare a weighted sum of the current through the source of one of the power transistors 402a' and 402b' and the input voltage with a reference voltage. The reference voltages are adjusted by two pairs of voltage dividing resistors 422a' and 424a', and resistors 422b' and 424b'. Preferably, the resistance of resistor 422a' equals the resistance of resistor 422b, and the resistance of resistor 424a' equals the resistance of resistor 424b', thereby causing the two reference voltages to be equal provided the resistors 424a' and 424b' are connected to the same voltage potential.

[0158] The resistors 426a', 428a' and 430a', and resistors 426b', 428b' and 430b', along with resistors 410a' and 410b', are scaling resistors. The relative values of the resistances of these resistors establishes the relationship between the input signal, the threshold currents and the reference voltage. The appropriate resistances for each of the resistors can be determined from simple principles of voltage division. In practice, it is preferable to have the resistance of resistor 410a' equal to the resistance of resistor 410b', the resistance of resistor 426a' equal to the resistance of resistor 426b', and the resistance of resistor 430a' equal to the resistance of resistor 430b', and the resistance of resistor 428a' equal to the resistance of resistor 428b', thereby assuring that each

comparator uses the same weighted sum of the input voltage, source current and reference voltage.

[0159] As described above, the structure of much of the amplifier circuit **400'** is symmetrical. That is, one of the halves processes the negative portion of the input signal and the other processes the positive portion. Inverter **700**, comprising of operational amplifier **702** and resistors **704** and **706**, inverts the input signal for the second branch of the circuit.

[0160] Resistors **444a'**, **446a'**, **444b'** and **446b'** are pull up resistors that are necessary for the proper operation of the voltage comparators **418a'** and **418b'**. Capacitors **448a'** and **448b'** provide a feedback loop for the current sensing comparators between the output and non-inverting input. This allows for faster switching times and short dead time zones, which is necessary for the proper functioning of the R-S flip-flop. Capacitors **450a'**, **450b'**, **452a'** and **452b'** are used to reduce high frequency noise and to improve circuit stability.

[0161] Transformer **404'** provides for energy storage and transfer. When one of the power transistors **402a'** or **402b'** is ON, energy is accumulated within the inductance of the transformer **404'** as a result of current flowing through one of the windings of the transformer **404'**. When the power transistor is turned OFF, and the other transistor is switched on, the energy accumulated in the inductance is released into the second winding as an electrical current.

[0162] It is important to achieve effective coupling between both transformer windings **404a'** and **404b'** in order to maximize efficiency and limit parasitic inductances that can lower the efficiency of energy transfer and cause voltage spikes in the drain of the switched OFF power transistor. If the voltage spike reaches the transistor breakdown voltage, then this additional energy is discharged through the transistor, dissipating additional heat. This heat energy dissipated should not be higher than the rated avalanche energy of the transistor. However, in high power applications, it is often difficult to limit this energy to below the avalanche energy of the transistor. To compensate for this, the amplifier **400'** includes a protective clamping circuit comprising clamping to diodes **454a'** and **454b'**, capacitors **456a'** and **456b'**, resistors **458a'** and **458b'**, discharge circuit, including transistor **460'**, resistor **462'**, capacitor **464'** and zener diodes **466'**, **468'** and **470'**. The clamping voltage should be less than the transistor breakdown voltage, and is selected by the zener diodes **466'**, **468'** and **470'**.

[0163] Various alternatives to the preferred embodiments of the amplifier are possible. For example, the amplifier could be implemented using a variety of technologies including, but not limited to, printed circuit board and integrated circuit implementations. In addition, the power switching transistors **72** and **74** of **FIG. 19** are indicated as being MOSFET transistors yet the design may be implemented using bipolar transistors as well. This variation may require other modifications as well, such as diodes connected between the emitters and collectors of these transistors; however, the scope of the invention would not be departed from. Furthermore, a number of aspects of the designs shown in **FIGS. 19 and 25** are, although useful, not integral to the invention. For example, the clamping networks and filtering capacitors may be omitted. If the filtering capacitors are omitted, then the high frequency currents i_1

and i_2 will be supplied to the coils rather than merely the average currents i_{a1} and i_{a2}. However, while this is not preferred, the speaker will still operate as the high frequency signals are at a sufficiently high frequency that a listener will only be able to hear the average audio output resulting from these high frequency signals. Additionally, although there are many advantages to controlling the power transistors based on the current sensing techniques described above, this is not necessary and alternative controlling techniques may be used. All such modifications or variations are believed to be within the sphere and scope of the invention as defined by the claims appended hereto.

1. A switching power amplifier for connection to a split voice coil of a speaker, the split voice coil having a first coil and a second coil located in a magnetic field, the first coil and the second coil being wired such that current can be provided to one coil without providing current to the other coil, the power amplifier comprising:

    a) a first branch for providing a first audio current to the first coil;

    b) a second branch for providing a second audio current to the second coil, wherein the first audio current differs from the second audio current, and

    c) a transformer for transferring energy between the first branch and the second branch such that both the first audio current and the second audio current constructively contribute to a force provided to the voice coil.

2. The switching power amplifier as defined in claim 1 wherein:

    the first branch comprises a first switching element for controlling a first current in the first branch;

    the second branch comprises a second switching element for controlling a second current in the second branch, wherein

        the first current and the second current have a selected high frequency period,

        the first audio current is the average of the first current over the high frequency period, and

        the second audio current is the average of the second current over the high frequency period.

3. The switching power amplifier as defined in claim 2 wherein the transformer comprises a first winding in the first branch and a second winding in the second branch, the first winding having a first terminal for receiving current flowing from the first branch and the second winding having a second terminal for receiving current flowing from the second branch, the first winding in the first branch and the second winding in the second branch being oriented such that the first terminal and the second terminal are of opposite polarity.

4. The switching power amplifier as defined in claim 2 wherein the transformer is one of a high frequency transformer and an impulse transformer.

5. The switching power amplifier as defined in claim 2 further comprising

    a first filtering capacitor for filtering high frequency components from the first current, the first filtering capacitor being connected to the first branch; and,

a second filtering capacitor for filtering high frequency fluctuations from the second current, the second filtering capacitor being connected to the second branch.

6. The switching power amplifier as defined in claim 2 wherein

the first audio current is provided concurrently with the second audio current, and

the first current and the second current are alternately supplied to the first branch and the second branch respectively such that the second current is interrupted when the first current is supplied to the first branch, and the first current is interrupted when the second current is supplied to the second branch.

7. The switching power amplifier as defined in claim 6 further comprising switching means for controlling

(i) the first switching element to terminate the first current at a variable first threshold magnitude at the end of a first time interval in each high frequency period,

(ii) the second switching element to terminate the second current at a variable second current termination magnitude at the end of a second time interval in each high frequency period; wherein

the transformer is a substantially 1:1 transformer, the first winding having a first terminal connected to the first switching element and the second winding having a second terminal connected to the second switching element, the second terminal being opposite in polarity to the first terminal such that the transformer is operable to transfer energy

from the first branch to the second branch during of the first interval to commence the second current at the associated first threshold magnitude, and

from the second branch to the first branch during the second interval to commence the first current at the associated second threshold magnitude.

8. The switching power amplifier as defined in claim 6 wherein the switching means is operable to control the variable first threshold magnitude and the variable second threshold magnitude from high frequency period to high frequency period to control the first audio current and the second audio current.

9. The switching power amplifier as defined in claim 2 further comprising

a power supply connection means for providing power to the first coil and the second coil; and,

a signal input branch for receiving an input signal providing audio information to the speaker; and

at least one signal circuit for dividing the input signal to provide the first signal and the second signal.

10. A method of providing an audio current comprising a first audio current and a second audio current to a split voice coil speaker having a first coil and a second coil located in a magnetic field, the first coil and the second coil being wired such that current can be provided to one coil without providing current to the other coil, the method comprising:

a) providing the first audio current to a first branch connected to the first coil;

b) providing the second audio current to a second branch connected to the second coil, wherein the first audio current differs from the second audio current;

c) orienting the first audio current in the first coil and the second audio current in the second coil such that both the first audio current and the second audio current constructively contribute to a force provided to the voice coil.

11. The method as defined in claim 10 wherein

step (a) comprises providing a first current to the first branch;

step (b) comprises providing a second current to the second branch;

the first current and the second current have a selected high frequency period,

the first audio current is the average of the first current over the high frequency period, and

the second audio current is the average of the second current over the high frequency period.

12. The method as defined in claim 11 further comprising transferring energy between the first branch and the second branch to make both the first current and the second current bi-directional.

13. The method as defined in claim 12 wherein the step of transferring energy between the first branch and the second branch is performed by a transformer having a first winding in the first branch and a second winding in the second branch.

14. The method as defined in claim 13 wherein the transformer is one of a high frequency transformer and an impulse transformer.

15. The method as defined in claim 12 further comprising

filtering high frequency components from the first current to provide the first audio current; and,

filtering high frequency components from the second current to provide the second audio current.

16. The method as defined in claim 12 wherein

the first audio current is provided concurrently with the second audio current, and

the first current and the second current are alternately supplied to the first branch and the second branch respectively such that the second current is interrupted when the first current is supplied to the first branch, and the first current is interrupted when the second current is supplied to the second branch.

17. The method as defined in claim 16 wherein

the first branch comprises a first switching element for controlling a first current; and,

the second branch comprises a second switching element for controlling a second current.

18. The method as defined in claim 17 further comprising

controlling the first switching element to terminate the first current at a variable first threshold magnitude at the end of a first time interval in each high frequency period;

controlling the second switching element to terminate the second current at a variable second current termination magnitude at the end of a second time interval in each high frequency period;

transferring energy from the first branch to the second branch during the first interval to commence the second current at the associated first threshold magnitude; and,

transferring energy from the second branch to the first branch during the second interval to commence the first current at the associated second threshold magnitude.

19. The method as defined in claim 18 further comprising controlling the variable first threshold magnitude and the variable second threshold magnitude from high frequency period to high frequency period to control the first audio current and the second audio current.

20. The method as defined in claim 11 further comprising dividing an input signal to provide the first signal and the second signal.

21. The method as defined in claim 11 wherein the first current and the second current are selected to be substantially inaudible when provided to the first coil and the second coil to impede sound distortion.

22. A speaker comprising:

a split voice coil having a first coil and a second coil located in a magnetic field, the first coil and the second coil being wired such that current can be provided to one coil without providing current to the other coil, and,

a power amplifier comprising

(a) a first branch for providing a first audio current to the first coil;

(b) a second branch for providing a second audio current to the second coil, wherein the first audio current differs from the second audio current, and

(c) a transformer for transferring energy between the first branch and the second branch such that both the first audio current and the second audio current constructively contribute to a force provided to the voice coil.

23. The speaker as defined in claim 22 wherein:

the first branch comprises a first switching element for controlling a first current in the first branch;

the second branch comprises a second switching element for controlling a second current in the second branch, wherein

the first current and the second current have a selected high frequency period,

the first audio current is the average of the first current over the high frequency period, and

the second audio current is the average of the second current over the high frequency period.

24. The speaker as defined in claim 23 wherein

the transformer comprises a first winding in the first branch and a second winding in the second branch, the first winding having a first terminal for receiving current flowing from the first branch and the second winding having a second terminal for receiving current flowing from the second branch, the first winding in the first branch and the second winding in the second

branch being oriented such that the first terminal and the second terminal are of opposite polarity.

25. The speaker as defined in claim 23 wherein the transformer is one of a high frequency transformer and an impulse transformer.

26. The speaker as defined in claim 23 wherein the power amplifier further comprises

a first filtering capacitor for filtering high frequency components from the first current, the first filtering capacitor being connected to the first branch; and,

a second filtering capacitor for filtering high frequency components from the second current, the second filtering capacitor being connected to the second branch.

27. The speaker as defined in claim 23 wherein

the first audio current is provided concurrently with the second audio current, and

the first current and the second current are alternately supplied to the first branch and the second branch respectively such that the second current is interrupted when the first current is supplied to the first branch, and the first current is interrupted when the second current is supplied to the second branch.

28. The speaker as defined in claim 27 wherein the power amplifier further comprises switching means for controlling

(i) the first switching element to terminate the first current at a variable first threshold magnitude at the end of a first time interval in each high frequency period,

(ii) the second switching element to terminate the second current at a variable second current termination magnitude at the end of a second time interval in each high frequency period; wherein

the transformer is a substantially 1:1 transformer, the first winding having a first terminal connected to the first switching element and the second winding having a second terminal connected to the second switching element, the second terminal being opposite in polarity to the first terminal such that the transformer is operable to transfer energy

from the first branch to the second branch during of the first interval to commence the second current at the associated first threshold magnitude, and

from the second branch to the first branch during the second interval to commence the first current at the associated second threshold magnitude.

29. The speaker as defined in claim 27 wherein the switching means is operable to control the variable first threshold magnitude and the variable second threshold magnitude from high frequency period to high frequency period to control the first audio current and the second audio current.

30. The speaker as defined in claim 22 further comprising

a power supply connection means for providing power to the first coil and the second coil; and,

a signal input branch for receiving an input signal for providing audio information to the speaker; and

at least one signal circuit for dividing the input signal to provide the first signal and the second signal.

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