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ACTIVE LOW PASS FILTER**TECHNICAL FIELD**

5 The present invention relates to an active low-pass filter.

BACKGROUND

10 In a digital cordless telephone, for example, a quadrature-type FM demodulation circuit is required, where a phase shifter circuit with a 90-degree input is used. This 90-degree phase shifter circuit generates a signal 90 degrees out of phase with respect to an input signal having a center frequency of f_0 as shown in Fig. 7; in this case, it is desirable that the its Q value (quality factor indicative of how sharp the peak shape of the

15 center frequency f_0 is) is as great as possible.

 Prior art examples of phase shifter circuit of this type are shown in Figs. 8-10. The circuit in Fig. 8 is a passive high-pass filter that is formed by combining a resistor R, a coil L, a capacitor C0 and a varicap diode CV, where the adjustment of the center frequency f_0 is made by changing the

20 terminal voltage of the varicap diode CV to vary its capacitance value.

 The circuit shown in Fig. 9 is a typical bi-quad low-pass filter formed by combining operational amplifiers A1-A3, resistors R1-R6 and capacitors C1 and C2. The circuit depicted in Fig. 10 is an active low-pass filter that has already been filed by the inventor of the present invention, where it is

25 comprised of three voltage-current amplifiers (transconductance amplifiers) A1-A3, resistors R1-R3, capacitors C1 and C2, and a current controller adj. This low-pass filter (LPF) is a second-order filter (as may become apparent later by solving the transfer function), such that its cut-off frequency f_0 is

30 adjusted by varying the amplification factors of the amplifiers A2 and A3 with the current controller adj. Exactly speaking, the center frequency and cut-off frequency have different meanings: however, in order to avoid complexity of symbols used, both are hereinbelow denoted by f_0 . It should also be noted that although the same symbols are used in the drawings, the same

35 symbols used in different drawings do not necessarily have the same characteristics.

 The passive high-pass filter illustrated in Fig. 8 is easier to adjust than those designed to adjust the center frequency f_0 by changing the coil

L; however, it requires adjusting the terminal voltage of the varicap diode, and it is difficult to reduce its size and weight because it is comprised of external parts. Even if variations in DC voltage of the FM demodulation output are used to control the terminal voltage of the varicap diode, thereby
5 eliminating the need for adjustment, its external circuitry become complicated to increase the size of the filter itself; in addition, as the temperature dependency of its external parts are exhibited, f_0 and Q become increasingly dependent upon temperature.

For the active LPF shown in Fig. 9, there are many external parts,
10 which requires larger real estate and results in higher cost. As such, the filters shown in Fig. 8 and 9, although used in portable products such as cordless telephone units, are not suited to implement reduction in size and weight and eliminate the need for adjustment.

On the other hand, with the bi-quad LPF depicted in Fig. 10, because
15 the amplification factors of the amplifiers A2 and A3 are changed by the current controller adj to adjust its cut-off frequency f_0 , it is superior to the afore-described circuits in that fewer external parts are required and that a reduction in size and elimination of adjustment can be implemented.

The present invention is intended to further improve the active LPF
20 shown in Fig. 10, and its purpose is to provide an active low-pass filter which allows for reduction in size and weight and can minimize its output noise.

SUMMARY OF THE INVENTION

25 The invention according to Claim 1 comprises an amplifier for acquiring a frequency signal into one input terminal and amplifying a differential voltage between one and other input terminals to output it as a current; a first capacitive element provided between the output terminal of the amplifier and a first potential source; a first impedance converter having
30 its input terminal connected via a first resistor to the output terminal of said amplifier; a second impedance converter having its input terminal connected via a second resistor to the output terminal of said first impedance converter and having its output terminal connected to the other input terminal of said amplifier; a second capacitive element connected
35 between said second resistor and the input terminal of said second impedance converter and the first potential source; and a third capacitive element having one terminal connected between said first resistor and the

input terminal of said first impedance converter, and having the other terminal connected to the output terminal of said second impedance converter.

5 The invention according to Claim 2 further comprises a third resistor connected between the other input terminal of said amplifier and a second potential source; and a fourth resistor having one terminal connected between the other input terminal of said amplifier and said third resistor, and having the other terminal connected to the output terminal of said second impedance converter.

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BRIEF DESCRIPTION OF THE DRAWINGS

Fig. 1 is a circuit diagram illustrating one example of how the present invention is implemented.

15 Fig. 2 is a circuit diagram illustrating one embodiment of the present invention.

Fig. 3 is a characteristics chart illustrating frequency and phase characteristics determined by simulation based on the active LPF of the present invention.

20 Fig. 4 is a characteristics chart illustrating frequency and phase characteristics determined by simulation based on the active LPF of the present invention.

Fig. 5 is a characteristics chart illustrating frequency and phase characteristics determined by simulation based on the active LPF of the present invention.

25 Fig. 6 is a characteristics chart illustrating frequency and phase characteristics determined by simulation based on the active LPF of the present invention.

Fig. 7 is a characteristics chart illustrating frequency and phase characteristics of an LPF.

30 Fig. 8 is a circuit diagram illustrating a prior art passive band-pass filter.

Fig. 9 is a circuit diagram illustrating a prior art active LPF.

35 Fig. 10 is a circuit diagram illustrating a comparative example of the active LPF.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENT

Fig. 1 is a circuit diagram illustrating one example of how the present invention is implemented. In Fig. 1, reference numeral 1 denotes an
 5 amplifier (transconductance amplifier) for acquiring a frequency signal into its positive input terminal and amplifying the input voltage to output it as a current.

To an output terminal of the transconductance amplifier 1 is connected one terminal of a capacitor C0, which comprises a first capacitive
 10 element, the other terminal of which is connected to a first potential source, e.g., a high potential source VCC. This high potential source VCC is at a +VCC potential. One terminal of the capacitor C0 is connected via a resistor R3, which comprises a first resistor element, to an input terminal of an emitter follower 2, which comprises a first impedance converter. To the
 15 output terminal of the emitter follower 2 is connected via a resistor R, which comprises a second resistor element, one terminal of a capacitor C1, which comprises a second capacitive element, while the other terminal of the capacitor C1 is connected to the high potential source VCC.

One terminal of the capacitor C1 (a node between the capacitor C1
 20 and resistor R) is connected to an input terminal of an emitter follower 3, which comprises a second impedance converter. The output terminal of the emitter follower 3 corresponds to an output terminal of the active LPF according to the present invention, such that this output terminal is connected via a resistor R2, which comprises a forth resistor element, to a
 25 negative input terminal of the transconductance amplifier 1. To a node between the negative input terminal of the transconductance amplifier 1 and the resistor R2 is connected one terminal of a resistor R1, which comprises a third resistor element, while the other terminal of the resistor R1 is AC grounded to earth, which is at a low potential. It should be noted that earth
 30 or ground, in this example, corresponds to a second potential source. Furthermore, the output terminal of the second emitter follower 3 is connected to the input terminal of the first emitter follower 2 via C2, which comprises a third capacitive element.

So configured, the active LPF operates as follows. When a
 35 frequency signal is applied to the positive input terminal of the transconductance amplifier 1, the transconductance amplifier 1 amplifies a difference between the voltage of input signal V_{in} and a feedback voltage

that is fed back from the output of the active LPF and divided by the resistors R1 and R2, and outputs it as a current. A voltage V1 is developed at one terminal of the capacitor C0 according to this current. The voltage V1 is then converted by the resistor R3 and capacitor C2 to a voltage V2, which is, in turn, outputted from the emitter follower 2. The output voltage V2 of the emitter follower 2 is converted by the resistor R and capacitor C1 to a voltage V3, which is then extracted as an output voltage Vout of the active LPF, and also fed back via the resistor R2 and capacitor C2 to the respective inputs of the transconductance amplifier 1 and first emitter follower 2.

Now, the above active LPF is described as a low-pass filter that comprises a 90 degree phase shifter circuit. First, the transfer function of this circuit is determined. It should be noted here that various symbols, such as Vin, C0, and R3, for the portions shown in Fig. 1 not only represent the corresponding symbols for such portions for the sake of convenience, but also denote their corresponding voltage value, capacitance, and resistance value, respectively, so equations described below are written using those symbols. Also note that the third capacitor C2 has its capacitance set at K•C0. In each equation, S denotes a complex variable ($S = j\omega$, where ω is a frequency); and gm denotes DI/DV (where DI denotes variations in collector current of a transistor that forms an amplifier, while DV denotes variations in base-to-emitter voltage).

By determining the relations for Vin, Vout, V1 and V2, the following equations (1) through (3) hold true.

25

[Equation 1]

$$(V_{in} - V_{out} * \frac{R1}{R1+R2}) * g_m = SC0V1 + (\frac{V1-V2}{R3}) \quad (1)$$

30

$$V_{out} = \frac{1}{1+SC1R} * V2 \quad (2)$$

$$\frac{V1-V2}{R3} = (V2-V_{out})SKC0 \quad (3)$$

From the above Eqs. (1)-(3), the following equation (4) holds true, so that the transfer function may be expressed as Eq. (5).

35

[Equation 2]

$$\begin{aligned}
 & (V_{in} - V_{out} * \frac{R_1}{R_1 + R_2}) * g_m \\
 5 \quad & = SC_0 V_1 + (V_2 - V_0) SKC_0 \\
 & = S * C_0 \{ (V_2 - V_{out}) SKC_0 R_3 + V_2 \} + (V_2 - V_{out}) SKC_0 \\
 & = S^2 KC_0^2 R_3 (V_2 - V_{out}) + SC_0 V_2 + SKC_0 (V_2 - V_{out}) \\
 & = S^2 KC_0^2 R_3 \{ (1 + SC_1 R) V_{out} - V_{out} \} \\
 & \quad + SC_0 (1 + SC_1 R) V_{out} + SKC_0 \{ (1 + SC_1 R) V_{out} - V_{out} \} \\
 10 \quad & = S^3 KC_0^2 C_1 R R_3 V_{out} + S^2 (K + 1) C_0 C_1 R V_{out} + SC_0 V_{out} \quad (4)
 \end{aligned}$$

$$\begin{aligned}
 \frac{V_{out}}{V_{in}} &= \frac{g_m}{S^3 KC_0^2 C_1 R R_3 + S^2 (K + 1) C_0 C_1 R + SC_0 + \frac{R_1 * g_m}{R_1 + R_2}} \\
 &= \frac{\frac{g_m}{(K + 1) C_0 C_1 R}}{\frac{K}{1 + K} * C_0 R_3 S^3 + S^2 + \frac{1}{(1 + K) C_1 R} * S + \frac{g_m}{(1 + K) C_0 C_1 R} (\frac{R_1}{R_1 + R_2})} \quad (5)
 \end{aligned}$$

15

Now assuming that the angular frequency in the cut-off frequency f_0 is represented by w_0 , then w_0 may be written as Eq. (6) in consideration of the above Eq. (5).

20

[Equation 3]

$$W_0 = \frac{1}{\sqrt{(1 + K) C_0 C_1 R (\frac{1}{g_m}) (\frac{R_1 + R_2}{R_1})}} \quad (6)$$

25

Also, given the above Eq. (5), Q (quality factor) may be expressed as Eq. (7) with $w = w_0$, where Q^* is a value of Q with $R_3 = 0$, and written as Eq. (8).

[Equation 4]

$$Q = \frac{W_o^2}{-AW_o^3 + \frac{W_o^2}{Q^*}} = \frac{1}{\frac{1}{Q^*}(1-A^*Q^*W_o)} = \frac{Q^*}{1-AQW_o} \quad (7)$$

5

$$Q^* = \frac{(1+K)C_1R}{\sqrt{C_0 \left(\frac{1}{g_m}\right) \left(\frac{R_1+R_2}{R_1}\right)}} \quad (8)$$

where

$$A = \frac{K}{(1+K)} C_0 R_3 = K C_0 C_1 R R_3 \left(\frac{W_o}{Q}\right)$$

10

Thus, using Eq. (7), Q is written as Eq. (9).

[Equation 5]

$$Q = \frac{Q^*}{1 - K C_0 C_1 R R_3 W_o^3} = \frac{Q^*}{1 - \frac{K R_3}{(1+K) \left(\frac{1}{g_m}\right) \left(\frac{R_1+R_2}{R_1}\right)}} \quad (9)$$

15

Now referring to Fig. 1, a first-order filter is implemented by the transconductance amplifier 1 and capacitor C₀; another first-order filter is implemented by the resistor R₃ and capacitor C₂; and still another first-order filter is implemented by the resistor R and capacitor C₁. Thus, the
20
afore-described active LPF comprises a third-order filter, which can be seen from the fact that a term S³ is contained in the transfer function represented by Eq. (5). The third-order filter with R₃ can have a greater value of Q as compared to a second-order filter with R₃ = 0; with the circuit shown in Fig.
25
1, in the cut-off frequency f₀ (angular frequency w₀), the third-order component implemented by R₃ and K•C₀ serves to enhance the value of Q by reducing the imaginary component.

If w = w₀, then s = jw₀; by substituting it into Eq. (5), the transfer function is expressed in second-order form as Eq. (10).

[Equation 6]

$$\frac{V_{out}}{V_{in}} = \frac{\left(\frac{R1+R2}{R1}\right) * H * (W_o)^2}{S^2 + \frac{W_o}{Q} * S + (W_o)^2} \quad (10)$$

5

The transfer function of the above Eq. (5), although written in a third-order equation, is in unique form; in other words, it can be seen that the active LPF of the present invention is a unique third-order filter. It is because R3 is provided that the active LPF acts as a unique third-order
10 filter. Now, since Eq. (7) holds true, the input/output phase difference is 90 degrees, as may be seen from Eqs. (11), (12) and (13) described later.

Additionally, as can be seen from Eq. (9), by varying R3, the input/output phase q (phase) assumes either 0, -90, or +90 degrees at w0. This may be expressed as Eqs. (11)-(13). In this way, R3 also serves to
15 adjust the phase as well.

[Equation 7]

$$\text{If } W=W_o \text{ and } R3 < \frac{(1+K) * \left(\frac{1}{g_m}\right) * \left(\frac{R1+R2}{R1}\right)}{K},$$

20 then q (phase) = -90 degrees (11)

$$\text{If } W=W_o \text{ and } R3 = \frac{(1+K) * \left(\frac{1}{g_m}\right) * \left(\frac{R1+R2}{R1}\right)}{K},$$

then q (phase) = 0 degrees (12)

$$\text{If } W=W_o \text{ and } R3 > \frac{(1+K) * \left(\frac{1}{g_m}\right) * \left(\frac{R1+R2}{R1}\right)}{K},$$

25 then q (phase) = +90 degrees (13)

Thus, according to the present invention, the value of Q can be set greater than 10, for example, as can be seen from the results of experiment
30 described later, so that the passband of the LPF can be narrowed. As compared to the circuit shown in Fig. 10 as the prior art example, the total

number of circuit elements is reduced; especially because only one amplifier is used, less noise is introduced. Thus, all things considered, low noise and high C/N can be secured as compared to the prior art active phase shifter circuit, while the S/N ratio can be substantially improved by
 5 employing the LPF of the present invention in the FM demodulation system.

The Q value and f_0 may be adjusted with the DC gain of the transconductance amplifier 1, capacitance ratio of the capacitors and resistors, although it is more advantageous to adjust the Q value and f_0 with the resistors R1, R2 and said DC gain, while reducing the capacitors. This
 10 is because the capacitive elements typically occupy a large area when incorporated into integrated circuitry. Thus, it is beneficial to use the resistors R1, R2 in the present invention. Absent the resistors R1 and R2, the value of $(R1 + R2)/R1$ would be equal to 1, which would cause the adjustment of f_0 and Q to become dependent upon C0 and C1, and large
 15 C0 and C1 would be required, and, additionally, because the output of the LPF would be directly fed back to the input of the transconductance amplifier 1, the output would be distorted if the dynamic range of the transconductance amplifier is narrow. It should be appreciated that although a technique for increasing R is available, it is not desirable to
 20 increase R because thermal noise is also increased.

Additionally, because the DC gain of the amplifier 1 can be varied with the resistors R1 and R2, the output level of the LPF can also be adjusted. Furthermore, because a feedback is implemented from the output terminal of the LPF to the two-stage input terminal by use of C2, the value Q
 25 can be enhanced by the capacitance ratio of $(1 + K)$.

The active LPF of the present invention also offers the following advantage, in addition to the afore-described benefits. That is, by employing a phase error correction loop for the FM demodulation system, the need for adjusting the Q value and f_0 can be completely eliminated.
 30 Furthermore, because a phase shifter circuit (coil and discriminator) that is essential to FM demodulation is incorporated into the IC, reductions in application cost, as well as size and weight, can be achieved, and unnecessary emission induced by any phase coil can be eliminated, thereby reducing the influence on other RF circuits.

35 With the afore-described active LPF, the temperature dependency of ω_0 and Q value can be theoretically reduced to zero by adjusting the temperature coefficient of g_m of the amplifier 1. That is, ω_0 and Q value can

be expressed as Eqs. (14) and (15), respectively, because the temperature characteristics $T_c(C_0)$ and $T_c(C_1)$ of C_0 and C_1 , respectively, are both approximate to zero.

5 [Equation 8]

$$\begin{aligned}
 W_o &= \frac{1}{\sqrt{(1+K)C_0C_1R\left(\frac{1}{g_m}\right)\left(\frac{R_1+R_2}{R_1}\right)}} \\
 &= \frac{1}{\sqrt{(1+K)C_0C_1\left(\frac{R_1+R_2}{R_1}\right)}} * \sqrt{\frac{g_m}{R}} \quad (14)
 \end{aligned}$$

10 (which is proportional to $\sqrt{\frac{g_m}{R}}$, and $T_c(C_0) = T_c(C_1) = 0$)

$$\begin{aligned}
 Q &= \frac{\sqrt{\frac{(1+K)C_1R}{C_0\left(\frac{1}{g_m}\right)\left(\frac{R_1+R_2}{R_1}\right)}}}{1 - \frac{KR_3}{(1+K)*\left(\frac{1}{g_m}\right)*\left(\frac{R_1+R_2}{R_1}\right)}} \\
 &= \frac{\sqrt{\frac{(1+K)C_1}{C_0\left(\frac{R_1+R_2}{R_1}\right)}} * \sqrt{Rg_m}}{1 - \frac{K}{(1+K)*\left(\frac{R_1+R_2}{R_1}\right)} * (R_3g_m)}
 \end{aligned}$$

15 which is proportional to $\frac{\sqrt{Rg_m}}{1 - \frac{K}{(1+K)*\left(\frac{R_1+R_2}{R_1}\right)} * (R_3g_m)}$ (15)

because $T_c(C_0) = T_c(C_1) = 0$

Differentiating w_0 and Q with temperature T yields Eqs. (16) and (17).
[Equation 9]

$$5 \quad \frac{\partial W_0}{\partial T} = \frac{\partial W_0}{\partial g_m} * \frac{\partial g_m}{\partial T} + \frac{\partial W_0}{\partial R} * \frac{\partial R}{\partial T}$$

$$= \frac{1}{2} \sqrt{\frac{1}{g_m R}} \frac{\partial g_m}{\partial T} - \frac{1}{2} \sqrt{\frac{g_m}{R^3}} \frac{\partial R}{\partial T}$$

$$= \frac{1}{2} \sqrt{\frac{g_m}{R}} \left(\frac{1}{g_m} * \frac{\partial g_m}{\partial T} \right) - \frac{1}{2} \sqrt{\frac{g_m}{R}} \left(\frac{1}{R} * \frac{\partial R}{\partial T} \right)$$

10

$$= \frac{1}{2} \sqrt{\frac{g_m}{R}} \left(\frac{1}{g_m} * \frac{\partial g_m}{\partial T} - \frac{1}{R} * \frac{\partial R}{\partial T} \right) \quad (16)$$

$$\frac{\partial Q}{\partial T} = \frac{\partial Q}{\partial g_m} * \frac{\partial g_m}{\partial T} + \frac{\partial Q}{\partial R} * \frac{\partial R}{\partial T} \mid T_c(R) = T_c(R_3), R_3 = A * R$$

15

$$= \frac{\frac{1}{2} \sqrt{\frac{R}{g_m}} \left\{ 1 - \frac{K(A * R * g_m)}{(1+K)(\frac{R_1+R_2}{R_1})} \right\} - \sqrt{R g_m} \left\{ \frac{K(A * R)}{(1+K)(\frac{R_1+R_2}{R_1})} \right\}}{\left\{ 1 - \frac{K(A * R * g_m)}{(1+K)(\frac{R_1+R_2}{R_1})} \right\}^2} * \frac{\partial g_m}{\partial T}$$

$$+ \frac{\frac{1}{2} \sqrt{\frac{g_m}{R}} \left\{ 1 - \frac{K(A * R * g_m)}{(1+K)(\frac{R_1+R_2}{R_1})} \right\} - \sqrt{R g_m} \left\{ \frac{K * A * g_m}{(1+K)(\frac{R_1+R_2}{R_1})} \right\}}{\left\{ 1 - \frac{K(A * R * g_m)}{(1+K)(\frac{R_1+R_2}{R_1})} \right\}^2} * \frac{\partial R}{\partial T}$$

$$= \frac{\frac{1}{2} \sqrt{R g_m} \left\{ 1 - \frac{K(A * R * g_m)}{(1+K)(\frac{R_1+R_2}{R_1})} \right\} - \sqrt{R g_m} \left\{ \frac{K(A * R * g_m)}{(1+K)(\frac{R_1+R_2}{R_1})} \right\}}{\left\{ 1 - \frac{K(A * R * g_m)}{(1+K)(\frac{R_1+R_2}{R_1})} \right\}^2}$$

20

$$* \left(\frac{1}{g_m} * \frac{\partial g_m}{\partial T} + \frac{1}{R} * \frac{\partial R}{\partial T} \right) \quad (17)$$

Thus, given Eq. (16), if the temperature coefficients for g_m and R are made equal to each other, then $[T_c(g_m) = T_c(R)]$ and temperature variations for w_0 are theoretically reduced to zero. In addition, if the absolute values of the temperature coefficients for g_m and R are made equal to each other in Eq. (17), and thus assuming that $T_c(g_m) = -T_c(R)$, then temperature variability for Q value are theoretically reduced to zero.

Next, an embodiment of the active LPF according to the present invention is illustrated in Fig. 2.

In this embodiment, an input signal V_{in} is applied to a transconductance amplifier 1 through an emitter follower 4 that is comprised of a transistor 41. The transconductance amplifier 1 includes differentially connected transistors (11, 14) and (12, 13), and current controllers 15 and 16. The voltage V_1 at the output of the transconductance amplifier is converted via resistor R_3 to a voltage V_2 , which is then outputted to an emitter of a transistor 22 of a first emitter follower 2 which is comprised of transistors 21 and 22 connected in two stages. A second emitter follower 3 is configured so that bases of transistors 31, 32, and 33 are connected to the output of the resistor R , while bases of transistors 34 and 35 are connected to emitters of the transistors 31 and 32, respectively, thereby separating the feedback output terminal from the output terminal of the LPF where the output signal V_{out} is extracted. The output terminals are separated in that manner in order to avoid the influence of the capacitor C_2 on the next-stage circuit of the LPF. Note that reference numerals 51-60 in Fig. 2 denote constant current sources. In the above example, capacitors C_0 and C_1 are connected to V_{CC} , while the resistor R_1 is AC grounded; however, the capacitors C_0 and C_1 may be AC grounded (first potential source), while the resistor R_1 may be connected to V_{CC} (second potential source) according to the present invention.

A simulation circuit was created in accordance with the circuit of Fig. 1, and computer simulation was performed to determine how the gain of the LPF and input/output phase deviation would vary as the value of the resistor R_3 was changed. Figs. 3-6 show graphs indicating the results, where R_3 is either 5 k Ω , 2.5 k Ω , 0.5 k Ω , and 0 Ω . Note that C_0 and C_1 are 15 pF and 40 pF, respectively; K is 5; R_1 , R_2 , and R are 1.5 k Ω , 2 k Ω , and 12 k Ω , respectively. As can be seen from the results, the value of Q changes as R_3 is varied, and the value of Q can be as high as 30, for example, as shown in Fig. 3

Accordingly, it should be appreciated that an active LPF of the present invention is not limited to use in an FM demodulation circuit, but may be used in place of an external coil.

Thus, an active LPF according to the present invention offers
5 advantages that it is small and lightweight, and the output noise can be reduced, while attaining a high Q value.

What is claimed is :

**CLAIMS**

1. An active low-pass filter comprising:
an amplifier for acquiring a frequency signal into one input terminal
5 and amplifying a differential voltage between one and other input terminals
to output it as a current;
a first capacitive element provided between the output terminal of the
amplifier and a first potential source;
a first impedance converter having its input terminal connected via a
10 first resistor to the output terminal of said amplifier;
a second impedance converter having its input terminal connected
via a second resistor to the output terminal of said first impedance converter
and having its output terminal connected to the other input terminal of said
amplifier;
15 a second capacitive element connected between said second
resistor and the input terminal of said second impedance converter and the
first potential source; and
a third capacitive element having one terminal connected between
said first resistor and the input terminal of said first impedance converter,
20 and having the other terminal connected to the output terminal of said
second impedance converter.
2. The active low-pass filter of claim 1, the filter further comprising:
a third resistor connected between the other input terminal of said
25 amplifier and a second potential source; and
a fourth resistor having one terminal connected between the other
input terminal of said amplifier and said third resistor, and having the other
terminal connected to the output terminal of said second impedance
converter.