Design of Low-Voltage Low-Power Analog Integrated Circuits for Hearing Aids

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ABSTRACT

This paper deals with the design of analog integrated circuits which are able to be operated with a single supply voltage down to 1.05 volts, i.e. the voltage from a zinc-air cell by the end of its lifetime, and consume minimal supply current. The practical circuits are mainly intended for use in hearing aids.

First some design strategies, especially suited for this area, are discussed. It is argued that there are good reasons to choose current instead of voltage as the information-carrying quantity. Next, the use of indirect feedback is shown to be favourable compared with traditional feedback methods. Finally, it is shown that the use of class-AB operated circuits fits well with operation in the current domain.

Next the suitability of various basic amplifier stages in a low-voltage/low-power environment are discussed. The traditionally most commonly employed stage, i.e. the differential pair, is shown to have inferior properties compared with other basic stages. A new "alternative" differential pair without these drawbacks is introduced.

Two systems and various circuits for hearing aids are discussed. Some (older) circuits were designed without the recently found design strategies and basic amplifier stages, whereas in other (new) circuits their merits were used and will be shown. The new circuits appear to have entirely different topologies than the traditional ones. The circuits discussed are:

- A preamplifier for electret microphones
- A 0.9V voltage regulator circuit
- Two controllable (pre-)amplifiers
- Two second-order high-pass filters

1. INTRODUCTION

Since about ten years the interest in special topics in electronics research has remarkably been changed. One of these topics is the area of *low-voltage/low-power* electronics. This is caused by the increasing importance of battery-fed electronic equipment, such as hand-carried telephones, pagers, whether or not implanted medical devices, e.g. *pacemakers* and *hearing aids*, and many other devices.

As hearing aids commonly contain a great variety of electronic circuits such as *amplifiers, filters, automatic gain controls, remote controls, voltage regulators*, etc.,

they are possibly the most suitable devices to apply and check the design considerations and the possibilities of low-voltage/low-power circuits.
First some items that will *not* be discussed. Although some existing hearing aids contain digital circuits, this paper deals only with *analog circuits*. The approach will be rather *synthetical* than *analytical*. Furthermore, no simulated and measured results are reported. They can be found in various papers listed in the *references*. Finally, almost all circuits only contain *bipolar* transistors, but this is mainly a matter of arbitrary choice. Most circuits can directly be designed in a CMOS process as well.

We start with the discussion of the overall boundary conditions and special design strategies, which are valid in a low-voltage/low-power environment. Then a selection of the most suitable basic amplifier stages and circuits is made. Further, two examples of complete hearing aid systems are dealt with, and finally several functional circuits, such as (whether or not controllable) amplifiers, filters, and a voltage regulator are discussed.

2. *OVERALL BOUNDARY CONDITIONS AND SPECIAL DESIGN STRATEGIES*

Overall boundary conditions

In hearing aids we find almost the only area of electronics, where the chip dimensions itself are sometimes of crucial importance. This is caused by the small dimensions of some of the complete devices: the smallest application is an ear-channel device for children. Generally, only chips smaller than 3x3 mm^2 are suitable for all occurring applications. Furthermore, in close connection to this demand, the number of external components must be restricted as much as possible. Other obvious demands are

- *extremely low power consumption*
- *low-voltage: the devices must essentially be able to operate at a single supply voltage down to 1.05V.*

To demonstrate the last item, the curve in Fig. 2-1 is characteristic: it shows a typical discharge curve of a zinc-air cell, i.e. the most commonly used and accepted battery for hearing aids. After a voltage of 1.05 volts is reached, it is discharged within a few minutes. Hence, it is useless to design the circuit for lower supply voltages, *unless batteries with lower cell voltages are developed in the future.*

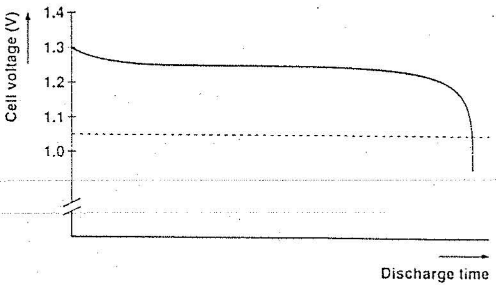


Fig. 2-1 Typical discharge curve of a zinc-air cell

Special design strategies

a) current level

By *current level* we mean a way of coupling between two devices (or circuits), so that the output impedance of the previous device (circuit) is very large compared with the input impedance of the following device (circuit). Consequently, the information carrying quantity is a *current*. There are a number of reasons why *current-level* should be chosen in our design constraints instead of *voltage*. These reasons are

- At current-level stages and/or circuits are better unilaterally coupled. At low power levels the output impedance of a device or circuit can attain very high values whereas the input impedance of the next stage (or circuit) remains relatively low, so that current-level is automatically attained.
- Multiple partition of the output signal of a circuit can easily be effectuated without introduction of large offset currents. (At voltage-level partition of signals needs more inputs and this is unattractive at low supply voltages and difficult realisable without the introduction of offset).
- At current level the parallel parasitics are short-circuited, i.e. their voltages become zero. At *low currents* these parasitics dominate over the series parasitics. Therefore current level is the best choice here.
- If the collector(drain) current of a device decreases, its input referred *voltage noise* increases whereas its input referred *current noise* decreases. Therefore, current coupling of circuits and/or devices is preferred.
- The application of Class (A)B-operated circuits is very favourable for the noise performance, if it is combined with current level, for if some signal changes from a large level down to its zero-crossing, the current noise decreases and the voltage noise increases. As the voltage noise has no effect in a coupling path that is completely at current-level, total input noise decreases with decreasing signal amplitude. This is very favourable in audio systems, because the audible noise is masked at high signal levels.

b) Indirect feedback

The most frequently method to improve the linearity and the accuracy of an amplifier is the use of *overall feedback*. Fig. 2-2a shows a well-known example of a current amplifier with *direct* overall feedback. However, in a low-voltage/low-power environment two problems arise. First, the voltage on the current-sensing resistor, i.e. the source of the MOST, does not fit with voltage at the input, hence a (difficult to realize) level shift is needed. Second, at very small *currents* the values of the resistors needed become very large (in practice sometimes a few giga-ohms), which cannot be realized on chip.

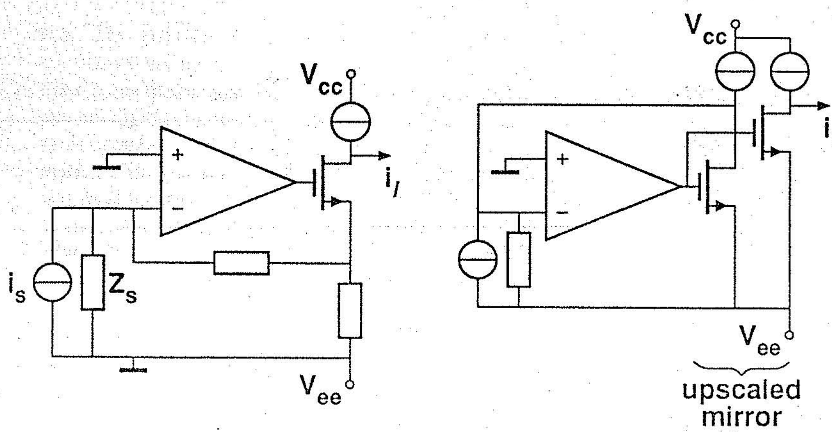


Fig. 2-2a,b. Current amplifiers with direct and indirect feedback.

A better method, where both problems are absent is shown in Fig. 2-2b. Now the drain current of the MOST is directly sensed and the gain is obtained by an upscaling current mirror. This method is called *indirect feedback.*

c) Multi-path frequency compensation

Apart from the traditional techniques for frequency compensation in overall feedback amplifiers, e.i. the insertion of *phantom zeros* and *pole-splitting techniques,* another method is sometimes useful in a low-voltage/low-power environment. It is based on lowering the order of an amplifier structure by implying first order parallel paths. Fig. 2-3 shows a simple example: The amplifying part of an overall feedback amplifier (A) which has two dominant poles is supplied with a first-order parallel path with lower gain and much larger bandwidth (B). After summing both transfers, the resulting transfer (C) has two dominant poles and one zero, which makes it a more stable system.

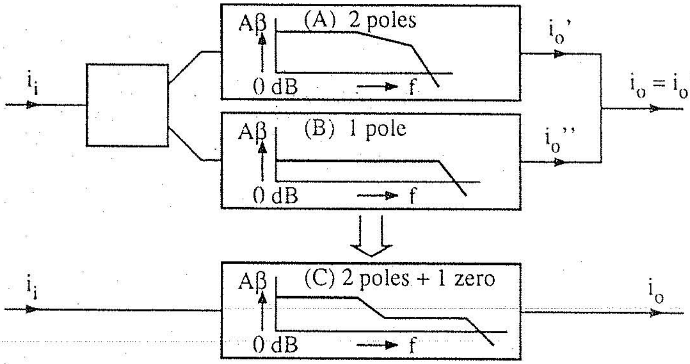


Fig. 2-3. Example of multi-path frequency compensation.

3. SELECTION OF THE MOST SUITABLE FUNCTIONAL STAGES AND BASIC CIRCUITS

Traditional analog circuits are generally teeming with long-tailed pairs (also called: differential pairs, DPs). However, this basic circuit (see Fig. 3-1) shows considerable drawbacks in a low-voltage/low-power environment, which are listed below.

- First, as only a *single* supply voltage down to 1.05V is available, a second voltage source is inevitable. Apart from the fact that its value is critical (it must be between about 0.75V and 0.90V), it must be low-ohmic and noise-free.
- Its input is typically at *voltage-level.* A conversion to current-level requires a (noisy) resistor, i.e. R in Fig. 3-1.
- As known its voltage-to-current transfer is very non-linear.
- Its quality "Common-Mode Rejection Ratio" (CMRR) is closely connected to operation at voltage-level. Hence it has little meaning as soon as circuits are current-level operated.

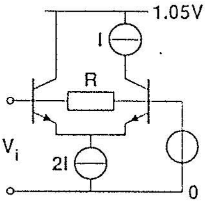
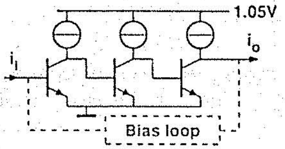


Fig. 3-1. Traditional differential pair.

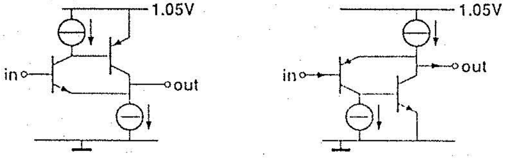
Better suited stages and basic circuits are:

- Cascaded common-emitter stages (CE-stages), see Fig. 3-2a;
- Two cascaded CE-stages, the two versions of which are shown in Figs. 3-2b,c;
- Controllable and/or scaled current mirrors, which have an *essentially* linear transfer [1];
- Basic translinear cells with up-down topology [2]; Fig. 3-3 shows an example of a multiplier/divider for simple mathematical manipulations with analog currents. The characters X,Y,U, and Z are (normalized) current values.

- CE-stage(s)

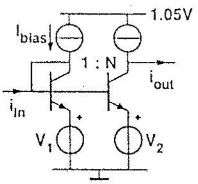


- 2 complementary CE-stages



- Controllable and/or scaled current mirrors

$$20\log\left|\frac{i_{out}}{i_{in}}\right| = 343(V_2-V_1) + 20\log N\ [\mathrm{dB}]$$



Figs. 3-a,b,c. Suitable basic amplifier stages in a low-voltage/low-power environment.

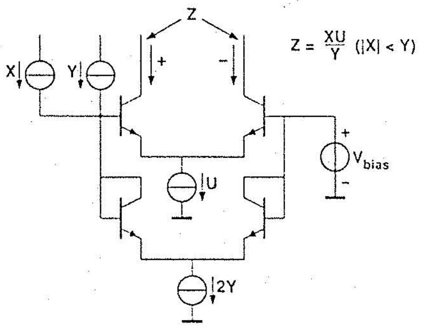
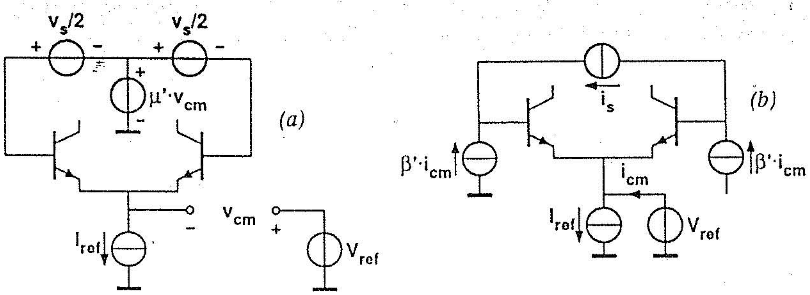


Fig. 3-3. Low-voltage translinear multiplier/divider.

Finally, we discuss a new differential stage, that is very suitable for current-level operation. It is called the "alternative" differential pair [3].



Figs. 3-4. Generalized traditional differential pair (a) and its dual form (b).

To come to its basic configuration, we start with a generalized form of the traditional DP (Fig. 3-4a). The only difference with a "normal" DP is that a the common-mode gain is enlarged with a voltage-controlled voltage source $\mu' \times Vcm$. With $\mu' = 0$ it degenerates to a normal DP. From Fig. 3-4a a dual form, completely operating at current-level, has been derived (Fig.3-4b). The duality appears from

- The chain matrices [4]: Let the chain matrix of a single CE stage be $\begin{pmatrix} A & B \\ C & D \end{pmatrix}$, then that of a traditional DP is $\begin{pmatrix} A & 2B \\ C/2 & D \end{pmatrix}$, and that of an alternative DP is $\begin{pmatrix} A/2 & B \\ C & 2D \end{pmatrix}$.
- The noise behavior: Let the input referred noise spectra of a single CE stage be $S(V)$ and $S(I)$, then those of the traditional DP are $2S(V)$ and $S(I)/2$, and those of the alternative DP are $S(V)/2$ and $2S(I)$.
- The common-mode behavior: A traditional DP has a large discrimination factor for *voltages* and an alternative DP for *currents* [3].

4. TWO EXAMPLES OF HEARING AID SYSTEMS

In this Section two hearing aid systems are discussed. We start with an infrared remote-controlled hearing aid system, depicted in Fig. 4.1. Infrared modulated code-words are received and demodulated by the IR-receiver and subsequently decoded by an I^2L-decoder whose outputs are 5 control bits for the volume control, 3 control bits for the high-pass filter control, and one stand-by control bit. The control bits are converted to analog control signals by D/A converters. The main analog string consists of a controlled preamplifier with two inputs, i.e. for microphone and telecoil, a high-pass filter, and a power amplifier. The remaining circuit blocks are for biasing arrangements. The sequence of preamplifier and filter has a great influence on the required dynamic range of the preamplifier: it must exceed the total dynamic range of the (microphone) input signals, i.e. >80dB.

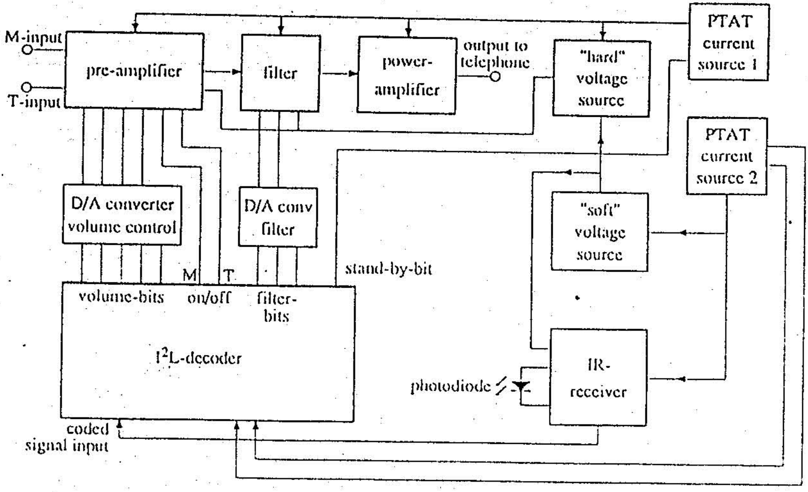


Fig. 4-1. IR remote-controlled hearing aid system.

Fig. 4-2 depicts a general purpose (but not remote-controlled) hearing aid system. As it contains all regularly occurring correction functions and it is suitable for all occurring output power levels, it can be used for almost all practical hearing aids. A feature is that all controls can be used by choice, i.e. if some control organ is not mounted, its control function changes into all-pass [5]. Below the simplest application is shown, i.e. a very small ear-channel device with only a volume control. Now the filters have been placed *before* the controllable amplifier. Consequently, the *filters* must have a dynamic range >80dB now! To realize this with integratable filter capacitors, the front-end is chosen to contain a companding system [5].

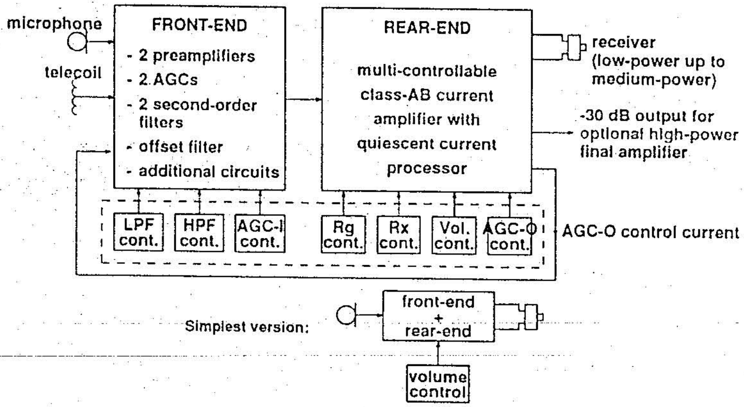


Fig. 4-2. General-purpose hearing aid system.

5. EXAMPLES OF HEARING AID CIRCUITS

The next circuits are briefly discussed: (for detailed data the referred papers should be consulted)

a)- A preamplifier for intrinsic electret microphones
b)- A 0.9V voltage regulator circuit
c)- Early voltage-level controllable preamplifier for standard electret microphones (and telecoils)
d)- Recent current-level controllable preamplifier for standard electret microphones
e)- Early voltage-level 2nd-order high-pass filter
f)- Recent current-level 2nd-order high-pass filter

Most of the circuits have been used in one of the systems shown in Figs. 4-1,2. In all examples one or more design principles such as they have been explained in Sections 1 through 3 can be recognized. In a few cases traditional solutions at voltage-level are compared with recent approaches at current-level.

a)- A preamplifier for intrinsic electret microphones [6]

Fig. 5-1 depicts the generic diagram. The electret is directly coupled to a completely balanced charge-to-current amplifier. As appears from the schematic with ideal biasing circuitry (Fig. 5-2), the common-mode biasing, effectuated by two diodes, carrying the gate leakage currents of the input MOSTs, is at current-level. This makes the structure that of an alternative differential pair, with its specific properties. Finally, Fig. 5-3 shows the complete circuit.

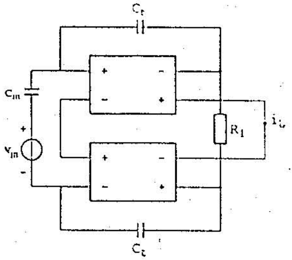


Fig. 5-1. Generic of the preamplifier.

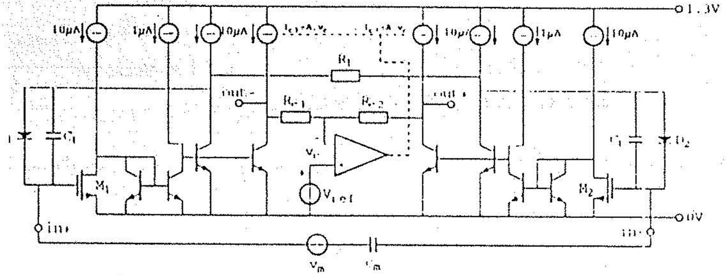


Fig. 5-2. Schematic with ideal biasing circuitry.

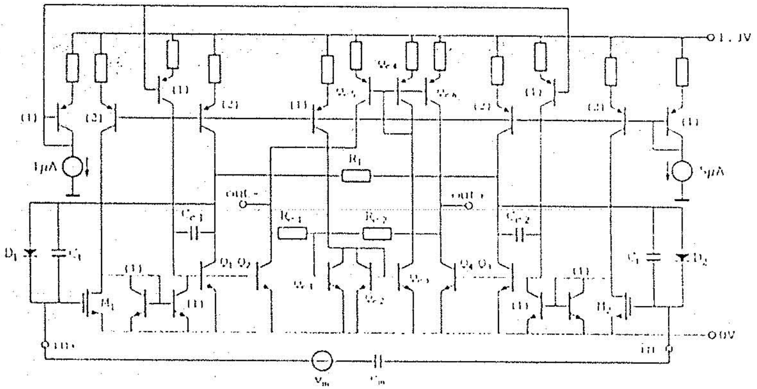


Fig. 5-3. Compete circuit.

b)- A 0.9V voltage regulator circuit [7]

Fig. 5-4 shows the block diagram of a traditional series regulator and Fig. 5-5 its current-level counterpart. The latter one is better suited for very low voltages, it can be designed with excellent noise properties, no external components are needed, and it effectuates automatic current limiting. The complete schematic is shown in Fig. 5-6.

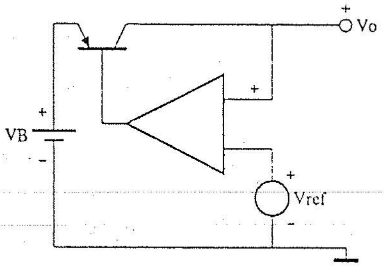


Fig. 5-4. Traditional series regulator.

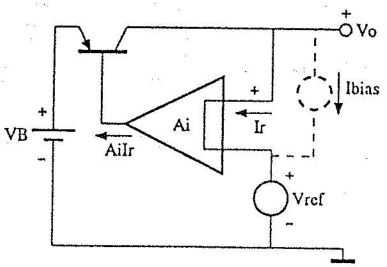


Fig. 5-5. Current-level series regulator.

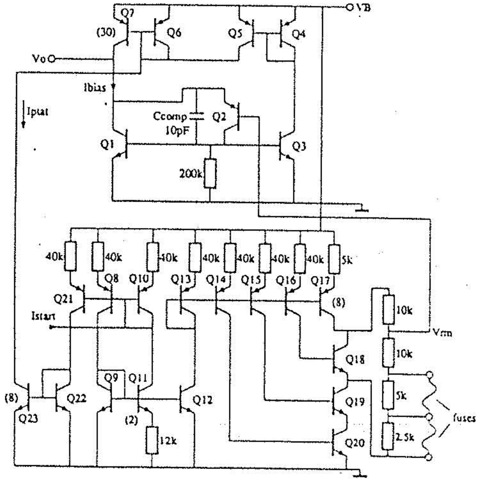


Fig. 5-6. Compete schematic.

c)- Early voltage-level controllable preamplifier for standard electret microphones (and telecoils) [8]

Fig. 5-7,8 depict the block diagram and the circuit diagram, respectively. It is part of the hearing aid system of Fig. 4-1. The gain control is effectuated by tail current control of three traditional differential pairs with all drawbacks, listed in Section 3. A current-level D/A converter provides for the correct tail currents.

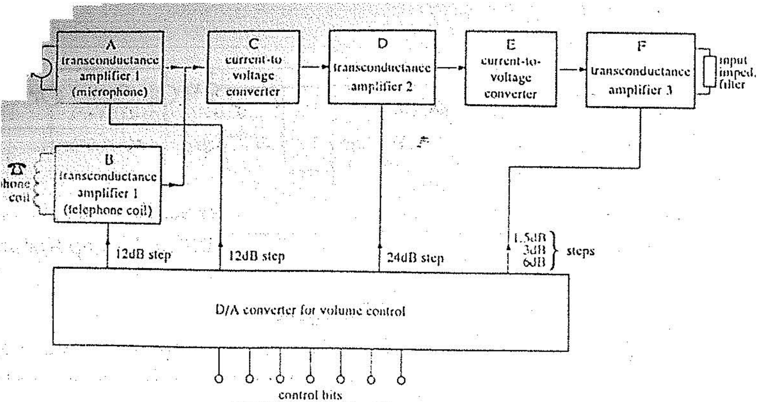


Fig. 5-7. Block diagram of an early controllable amplifier.

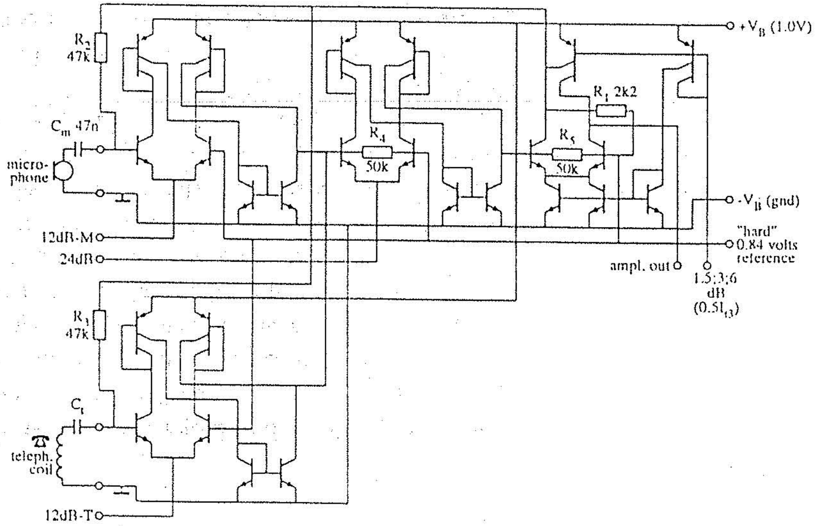


Fig. 5-8. Complete circuit diagram.

d)- Recent current-level controllable preamplifier for standard electret microphones [9]

Fig. 5-9 shows the basic circuit configuration. Now the use of differential pairs has been avoided and replaced by a controllable current mirror as described in Section 3. As a result, this circuit has better noise and efficiency properties than the previous one. As the circuit produces offset due to finite base currents and mismatch, an offset reducing feedback loop has been added (Fig. 5-10). The complete circuit is shown in Fig. 5-11.

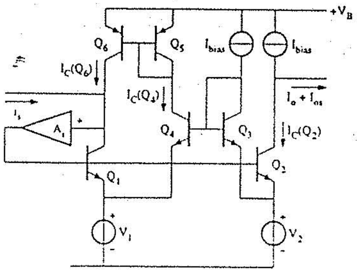


Fig. 5-9. Basic configuration of a recent controllable amplifier.

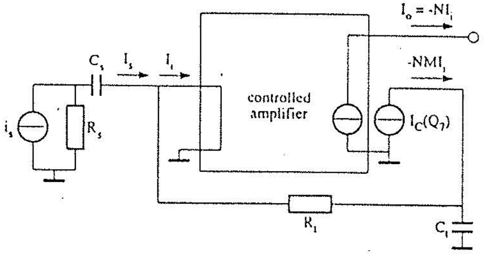


Fig. 5-10. Offset reducing feedback loop.

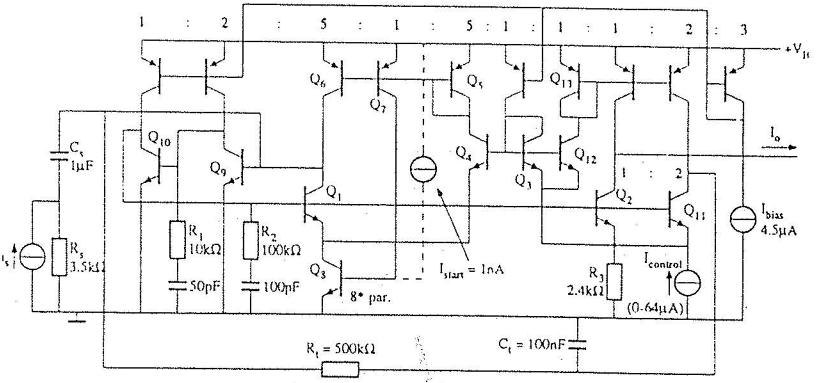


Fig. 5-11. Complete circuit.

e)- Early voltage-level 2nd-order high-pass filter [8]

The circuit is part of the system in Fig. 4-1. Figs. 5-12,13 show the block diagram and the circuit diagram, respectively. The filter is built up by two capacitor-loaded gyrators, coupled by a voltage-to-current converter. It provides two equal poles, whose frequency can be controlled by the tail currents of four traditional differential pairs. All drawbacks, mentioned in Section 3 are present.

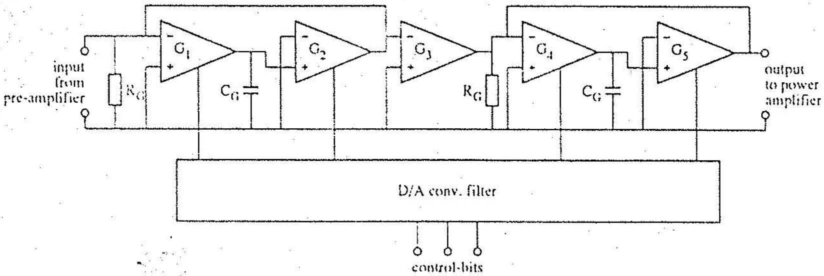


Fig. 5-12. Block diagram of an early second-order high-pass filter.

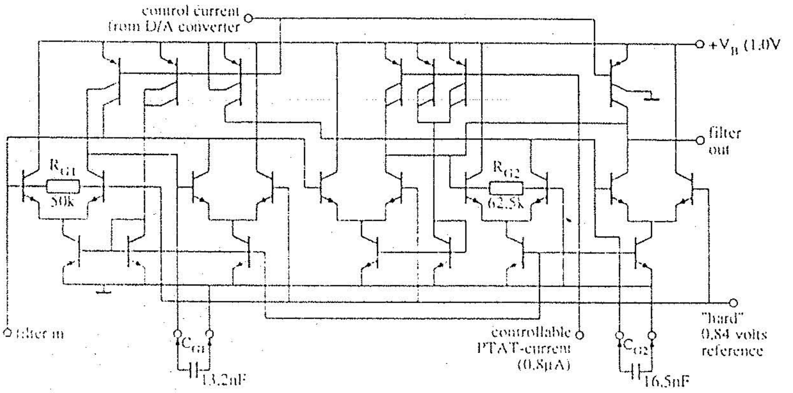
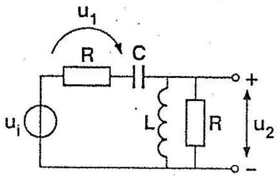


Fig. 5-13. Circuit diagram of the filter.

f)- Recent current-level 2nd-order high-pass filter [10]

This filter has been designed in a more systematic way. First an electric analogon with passive components together with its network equations has been used (Fig. 5-14). Then, a current-level version is obtained by replacing u_i, u_1, and u_2 by i_i, i_1, and i_2. Fig. 5-15 shows a leapfrog version of the filter, satisfying the resulting current equations. The lower blocks are integrators and the upper blocks are simply current mirrors. Fig. 5-16 depicts one of the employed integrators, where an npn-pnp combination is used as the active part (see Section 3). Finally, Figs. 5-17,18 show the resulting signal path and the compete circuit diagram. The corner frequency is controlled by the (continuous controllable) scaling factor of the output mirrors of the integrators.



$$\frac{u_1}{SA_1/\omega_c} = u_i - u_1 - u_2$$

$$\frac{u_2}{SA_2/\omega_c} = u_1 - u_2$$

Butterworth:

$A_1 = A_2 = \sqrt{2}$

$L_1 = RA_2/\omega_c$

$C = A_1/\omega_c R$

Fig. 5-14. Passive analogon with network equations of the filter.

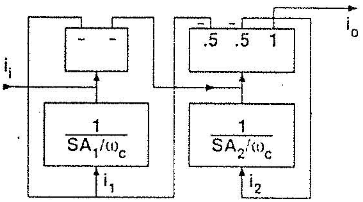


Fig. 5-15. Current-level leapfrog version of the filter.

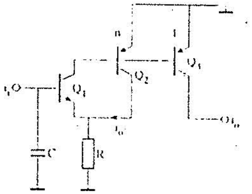


Fig. 5-16. One of the employed integrators.

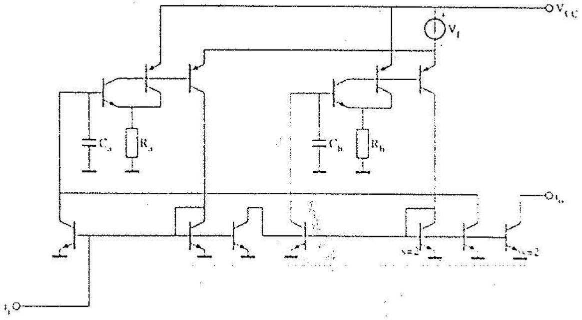


Fig. 5-17. Resulting signal path of the filter.

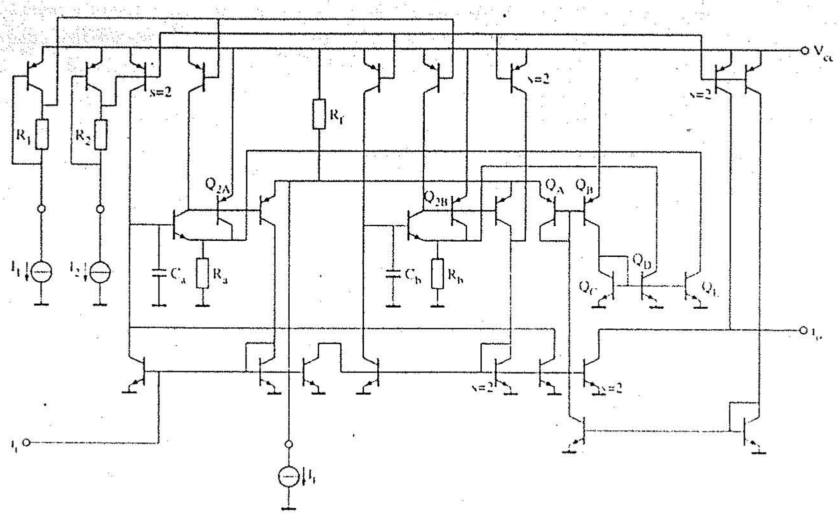


Fig. 5-18. Complete circuit diagram.

6 General conclusions

In this paper a number of general valid rules for the design of analog circuits operating at very low voltages and powers have been presented. These rules and some other conclusions can be summarized as follows:

- Avoid traditional differential pairs as much as possible
- Choose for circuits operating at current-level instead of at voltage-level
- Use translinear circuits for simple analog mathematical operations
- Combine operation at current-level with class-(A)B operation
- Use device matching as much as possible
- Low supply voltage (down to 1.05V) has never appeared to be a serious limitation for high-quality designs.

6 References

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