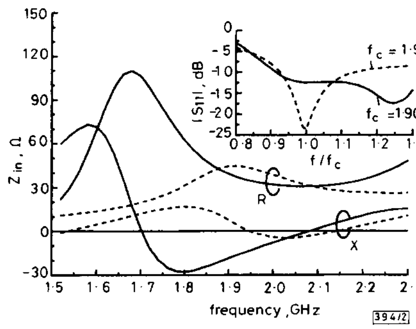

Fig. 2 *Impedance against frequency and return loss against normalised frequency for TM_{01} mode*

——— without gap
– – – – with gap

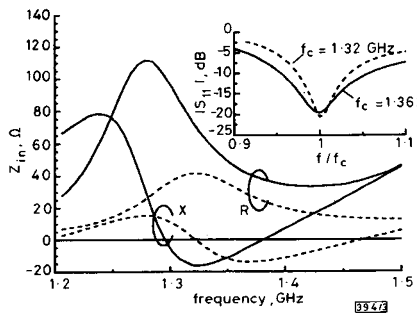

Fig. 3 *Impedance against frequency and return loss against normalised frequency for HEM_{11} mode*

——— without gap
– – – – with gap

slightly decreases the 10dB bandwidth, but does not improve the 50Ω match. The frequency at which the return loss is minimum is near that of the predicted resonance frequency of the CDR without the presence of the feed probe as shown in the return loss insertion in Fig. 3. Though not as obvious, the antenna with the air gap increases the resonance frequency of the next higher order mode. This effect can be seen by observing that R_{in} begins to increase near 1.425GHz for the no gap antenna, whereas it continues to decrease for the antenna with the air gap. Other length feed probes were used to excite these two antenna modes and the results of those experiments were similar.

Conclusions: For DR antennas with coaxial probe feeds, an air gap between the feed probe and the dielectric material can significantly affect the input impedance, resonance frequency, and bandwidth of the antenna. The effect may be especially severe for DR antennas composed of high permittivity materials and/or operating in the millimetre wave range. It may be possible to improve the performance of a DR antenna by using a coaxial probe coated with low permittivity material.

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7 December 1993

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Low-voltage, low-power, wide-range controllable current amplifier for hearing aids

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Indexing terms: Audio-frequency amplifiers, Circuit design

A controllable current amplifier with a control range of more than 60dB for application in a novel, completely integratable hearing aid is presented. It operates on power supply voltages of 1 – 1.3V. Low current consumption is aimed at. The maximum value is 93µA.

Introduction: A controllable current amplifier has been designed for a new, completely integratable hearing aid, the block diagram of which is given in Fig. 1.

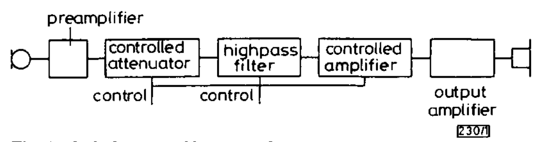

Fig. 1 *Block diagram of hearing aid*

Many users of hearing aids need the ability to improve speech intelligibility in environments with much low-frequency noise (e.g. cars). This can be achieved with the controllable highpass filter presented in [1]. To make this filter integratable, it has to operate at a very low current level (signal level 25nA peak). As a consequence, the signal-to-noise ratio is limited. To obtain a sufficiently large dynamic range, a controllable attenuator [2] and a controllable amplifier should keep the signal level in the filter at the largest possible value, independently of sound volume. This Letter presents such a controllable current amplifier. From the defined signal levels at the input and the output it follows that current gain has to be controllable between 0 and 60dB. The circuit has to operate at a very low supply voltage of 1 – 1.3V. In-line with usual specifications, the minimal required signal-to-noise ratio has been set at 50dB and the harmonic distortion at less than 5%. Furthermore the amplifier is optimised with regard to power consumption.

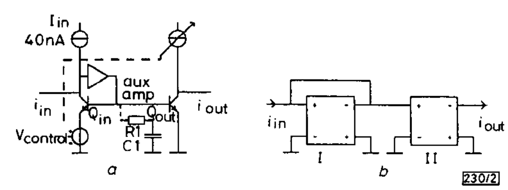

Fig. 2 *Controllable current mirror and indirect feedback*

a Controllable current mirror
b Indirect feedback

Basic configuration: The basic configuration is the controllable current mirror shown in Fig. 2*a*. In this circuit the influence of the base currents is reduced by means of an auxiliary amplifier.

The circuit is of the class of indirect feedback amplifiers [3], the principle of which is given in Fig. 2*b*. The inputs of the amplifier stages are placed in parallel. The output of the first stage is used for feedback, and the output of the second stage is used for driving the load. When the stages are identical, the current gain is −1. The advantages of this configuration are obvious: feedback can be applied without the need for sensing the current through the load. Moreover, in the controllable current mirror, the current gain can easily be made controllable by means of the voltage source $V_{control}$. When base currents are neglected, the current gain is given by

$$a_i = i_{out}/i_{in} = \exp(V_{control}/V_T) \qquad (1)$$

where V_T is the thermal voltage. Gain varies exponentially with the control voltage, hence a gain control in dB can be made easily: at room temperature the gain changes 20dB per 60mV. $V_{control}$ must be proportional to the absolute temperature (PTAT), which can easily be realised. The bias current of Q_{in} is chosen somewhat higher than the signal amplitude to avoid clipping. A value of 40nA has been chosen. Because the bias current only needs to be slightly larger than the signal amplitude, the circuit offers very good power efficiency. When the control voltage is increased, the DC current through Q_{out} increases exponentially so an output bias source which is controlled by $V_{control}$ is necessary.

Noise properties: Because of the auxiliary amplifier, base currents are negligible and the only relevant noise sources are the two collector shot noise currents of the transistors. When the noise source of Q_{out} is transformed to the input [4], the equivalent input noise current spectrum $S(i_{n,eq})$ is given by

$$S(i_{n,eq}) = 2q(1 + 1/a_i)I_{in} \qquad [A^2/Hz] \qquad (2)$$

Clearly, the dynamic range of the circuit hardly depends on the adjusted gain. This is an important property in this application. In practice, with a signal bandwidth of 10kHz, a signal-to-noise ratio of 57dB can be obtained.

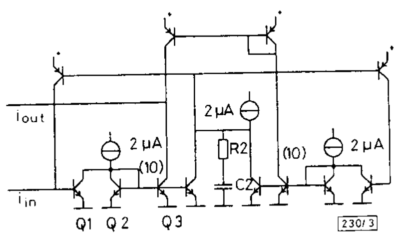

Fig. 3 *Auxiliary-amplifier circuit*

Auxiliary amplifier: The auxiliary amplilier has to enlarge the loop gain to such an extent that the base currents are negligibly small compared to the collector current of Q_{in}. The circuit is given in Fig. 3.

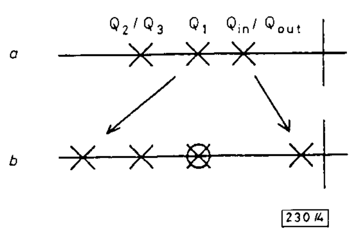

Fig. 4 *Pole positions in signal loop*

a Before compensation
b After compensation with R1/C1

This is a variant of the 'alternatively biased differential pair' [5]. The amplifier consists of two stages. The first stage, Q_1, is a CE-stage and the second stage, Q_2/Q_3, is an amplifying current mirror. The circuit is implemented symmetrically because then the input bias currents can be generated by means of a common-mode loop. Because of the load of the amplifier, with a *pnp* current mirror the common-mode loop has hardly any influence on the signal path.

The only condition is that it has to be stable, which is provided for by pole-zero cancellation (R_2/C_2) [3]. Now in the signal loop of Fig. 2, three poles occur (Fig. 4*a*).

Because the two poles of the auxiliary amplifier are far apart, compensation with a small, integratable capacitor is sufficient to obtain a stable loop. By using pole-zero cancellation again (R_1/C_1 in Fig. 2), the pole contributed by Q_{in}/Q_{out} is shifted to lower frequencies, while the pole that originates with Q_1 is apparently shifted towards a much higher frequency (Fig. 4). As gain is adjusted at increasingly higher values, the pole at the lowest frequency shifts towards higher frequencies and loop gain decreases, both because of the decreasing value of $r_{pi}(Q_{out})$. As a result, the positions of the closed loop poles hardly change, so the bandwidth remains the same.

The noise contribution of the auxiliary amplifier can be made small compared to the noise calculated in eqn. 2 by ensuring that $I_b(Q_1) << I_c(Q_{in})$. In this case the base current shot noise of Q_1 is negligibly small compared to the collector shot noise of Q_{in}, and eqn. 2 remains valid.

Biasing of controllable current mirror: Because the DC collector current of Q_{out} varies with gain adjustment, a variable output bias current source that exactly tracks the gain control has to be designed. This source is realised with a second current mirror which is also controlled by the source $V_{control}$ (Fig. 5). This current mirror has been extended with an additional amplifier as well. This amplifier can be much simpler because it is not a part of the signal path. As the impedance at the emitter of Q_{in} is relatively high, no high demands are made on the voltage source. The circuit was tested with a 30kΩ resistor which was biased with a controllable current. The noise contribution of the voltage source is usually negligible because it appears in the circuit as a common-mode effect.

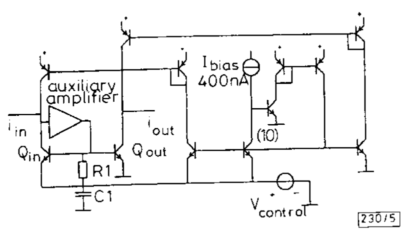

Fig. 5 *Biasing of controllable current mirror*

Realisation: The circuit has been realised in a Philips BiCMOS process in a semi-custom version. Of course, when the hearing aid is integrated on a single chip, a fully custom redesign will be made to save chip area. The values of the resistances and capacitances are: R_1 = 19kΩ, C_1 = 300pF, R_2 = 8kΩ, C_2 = 120pF. Fig. 6 shows the frequency behaviour of the circuit. The bandwidth is approximately 150kHz. The irregularity around ~15kHz is caused by some overshoot in the common-mode loop of the auxiliary amplifier. This effect can be overcome by using a larger compensation capacitor C_2, but this is not necessary. Moreover, a large capacitor is undesirable.

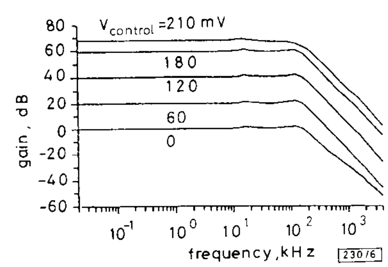

Fig. 6 *Current gain against frequency*

Conclusion: The proposed circuit is very simple but nevertheless provides an extremely large control range of more than 60dB, while current consumption remains very low. The signal-to-noise ratio is 57dB. THD increases with gain but remains less than 0.9%. Current consumption varies between 29 and 93μA.

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Minimum shunt capacitance of class E amplifier with optimum performance

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Indexing terms: Power amplifiers, Circuits design

The minimum value of the shunt capacitance of a class E amplifier is derived and is expressed in terms of the specified parameters of the amplifier. If the shunt capacitance is selected to be larger than this minimum value, the output power of the amplifier cannot reach the desired level. The effect of the shunt capacitance and the choke inductor on the output performance of class E amplifier is also illustrated.

Introduction:: The class E amplifier is a high-efficiency amplifier and its collector efficiency can reach as high as 100% if the optimum conditions are satisfied [1–3]. An analytical method was derived in [4] to provide a fast approach for evaluating the required component values of the class E amplifier to obtain optimum performance with the desired specifications. It is suggested in [4] that the value of the shunt capacitance can be selected to be equal to C_{ob}, the output intrinsic capacitance of the transistor, but it may not be applicable in certain applications, especially for low-frequency operation, because C_{ob} may be too small.

In this Letter, we show that a maximum value of the output power of the class E amplifier in optimum operation can be found if the required specifications and the shunt capacitance are fixed. If the desired output power of the amplifier is substituted into the expression for maximum output power, the calculated value of the shunt capacitance will correspond to the minimum value of that power output level. To obtain the desired specifications of the amplifier, the shunt capacitance must be selected greater than or equal to this minimum value.

Theory: According to the analytical method derived in [4], the output power P_{out} of the class E amplifier in optimum performance can be expressed in terms of the choke inductor L_1, the shunt capacitance C_1, and the operation frequency f; i.e.

$$P_{out} = V_{cc}\left[\frac{LA}{2\pi}\sin\frac{\pi}{L} + \frac{LB}{2\pi}\left(1-\cos\frac{\pi}{L}\right) + \frac{I_o\cos\varphi}{(1-L^2)\pi} + \frac{V_{cc}\pi}{L_1 4\omega} - \frac{I_o}{2}\sin\varphi\right] \quad (1)$$

where

$$L = \frac{\omega}{\sqrt{L_1C_1}} \quad (2)$$

$$A = \left(\frac{1}{1-\cos\frac{\pi}{L}}\right)\left[\left(\frac{V_{cc}}{L_1\omega_o} - \frac{I_oL}{1-L^2}\cos\varphi\right)\sin\frac{\pi}{L} - \frac{2I_o}{1-L^2}\sin\varphi + \frac{V_{cc}}{L_1}\left(\frac{\pi}{\omega}\right)\right] \quad (3)$$

$$B = \frac{V_{cc}}{L_1\omega_o} - \frac{I_oL}{1-L^2}\cos\varphi \quad (4)$$

$$I_o = \frac{V_{cc}(1-\cos\frac{\pi}{L} + \frac{\pi}{2L}\sin\frac{\pi}{L})(1-L^2)}{L_1\omega_o\sin\varphi\sin\frac{\pi}{L}} \quad (5)$$

$$\cot\varphi = \frac{\pi\omega_o\cos\frac{\pi}{L} + \omega\sin\frac{\pi}{L}}{\omega L(1-\cos\frac{\pi}{L} + \frac{\pi}{2L}\sin\frac{\pi}{L})} - \frac{2}{L}\cot\frac{\pi}{L} - \frac{L(1-\cos\frac{\pi}{L})}{\sin\frac{\pi}{L}} \quad (6)$$

A detailed derivation of these expressions can be found in [4].

When the value of L_1 approaches infinity, eqn. 1 with a fixed C_1 and f will tend to a limit and this limit is shown in eqn. 7:

$$\lim_{L_1\to\infty} P_{out} = 19.74C_1fV_{cc}^2 = P_{omax} \quad (7)$$

P_{omax} is the maximum output power of the class E amplifier. We may rearrange eqn. 7 as follows:

$$C_1 = \frac{P_{omax}}{19.74fV_{cc}^2} = C_{1min} \quad (8)$$

C_1 in eqn. 3 is the minimum required value of the shunt capacitance, C_{1min}. This means that if the shunt capacitance is selected lower than C_{1min}, the output power of the amplifier in this case cannot reach the desired output level. Consequently, in the design of a class E amplifier, we may calculate C_{1min} by eqn. 8 first and then select the shunt capacitance which is greater than the calculated C_{1min}.

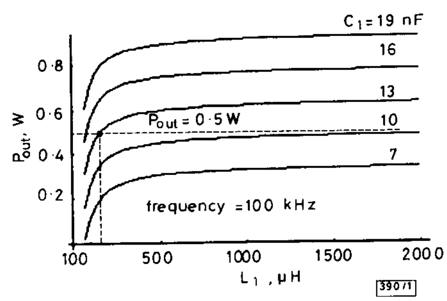

Fig. 1 *P_{out} against L_1 with parameter C_1*

Experimental results: For ease of demonstration in the time domain, a 100kHz class E amplifier was designed and implemented. The desired output power and operation voltage of the amplifier are 0.5W and 5V, respectively. C_{1min} is equal to 10nF from eqn. 8. We may plot a graph of P_{out} against L_1 with parameter C_1 in Fig. 1 according to eqn. 1. The shunt capacitance C_1 is selected to be 13nF from Fig. 1 and the required values of the other components can be calculated, and are shown in the first column of Table 1. A class E amplifier was then implemented and a 'Motorola' RF BJT transistor MRF559 used as the active device. The experimental results of the amplifier are shown in the second column of Table 1. In addition, the experimental waveforms of the collector voltage and current are shown in Fig. 2. From Fig. 2, we find that the amplifier has optimum performance, and the collector efficiency reached 93%.

When shunt capacitance C_1 is changed to 8nF (< C_{1min}), the input and output power is decreased. For the amplifier tuned to the best performance, the experimental values of components and output performance are given in the third column of Table 1. The waveforms of the collector voltage and current are shown in Fig. 3