Design of High-Performance Negative Feedback Amplifiers

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Preface

Amplifier design is very often regarded as making a selection from the large arsenal of known amplifier circuits and then adapting it for a specific purpose, possibly with the aid of computer-aided-design programs. Now and then designers are surprised by the introduction of a new amplifier circuit performing better in some respect than the others.

Each aspiring designer has to find his own way in this jungle. He has to choose from a rather chaotic and scattered collection of amplifier circuits rather than apply a systematic and straightforward design sequence that enables him to design his special-purpose amplifier circuit. A great deal of experience is essential.

This work is an attempt to make a useful contribution to the extensive literature on the subject of amplifier design. It can be justified on the grounds that the approach is believed to be unique in a number of respects. Many works that promise to cover the subject are instead concerned with analysis. Moreover, they frequently deal specifically or separately with particular design aspects, characterised by descriptions such as ‘wide-band’, ‘low-noise’, ‘low-distortion’, etc.

A treatment of the various design aspects and their interconnections, however, is necessary for fruitful amplifier design. At the basis of such a treatment lies the observation — usually easily overlooked — that amplifier design is concerned in the first place with obtaining an adequate quality of information transfer. Amplifiers are more than electronic circuits merely bringing the source power up to a higher level.

Quality requirements are imposed on the signal transfer relative to the type of information and to the manner of perception, registration, or processing. The quality of information transfer is determined by a large number of quality aspects such as linearity, accuracy, efficiency, signal-to-noise ratio, etc. Unfortunately, it cannot be expressed as a quantitative figure of merit

Trade-offs between various quality aspects are likely to emerge. Sometimes they will be of a fundamental nature and are imposed by physical and technological limitations, but frequently they will result from the nature of a specific amplifier circuit itself, which — on second thoughts — might not be the most appropriate one to fulfil the desired function.

A systematic, straightforward design approach is presented in this work. It is more or less inspired by the work of Cherry and Hooper[1] which I consider one of the finest design treatises. The present work is more concerned with basic design considerations. Preference was given to a qualitative rather than to a quantitative approach.

Finding the proper configurations for the basic amplifier and of the amplifier stages is considered of primary importance and is emphasised here. This book is therefore largely concerned with the design phase preceding the phase in which existing computer aids can be helpful.

The approach is characterised best by describing it as a systematic and consistent arrangement of design considerations regarding various quality aspects of information transfer. Via the classifications of amplifier configurations, a systematic design method for negative-feedback amplifiers is developed.

A short description of the main lines along which the design method has been developed is given below.

In chapter 1, criteria are given for the adaptation of the input and output impedances to the source and the load, respectively (usually transducers). The purpose of these adaptations is the realisation of optimum information transfer from the signal source to the amplifier and from the amplifier to the load. Next, criteria are deduced for optimum information transfer of the amplifier, preserving signal-to-noise ratio and efficiency by the application of feedback. Classifications are given of basic amplifier configurations with up to four negative-feedback loops, providing the designer with the complete set of fundamentally different two-port amplifier types. The characteristic properties and the practical merits of these configurations are discussed.

A similar classification is given in chapter 2 for configurations with a single active device. A uniform description of these single-device configurations will appear to be of great help in finding suitable stage configurations in the active part of an amplifier with overall negative feedback. The balanced versions of these single-device stages are mentioned but not studied in detail.

In chapter 3, design criteria regarding random noise are formulated. These criteria relate mainly to the selection of the most favourable input-stage configuration and the active device to be used in this stage for a given signal source.

In chapter 4, those configurations of amplifier stages that are best suited to realising optimum accuracy and linearity of information transfer are arrived at.

Thereafter, bandwidth and stability considerations are taken into account in chapter

[1] E.M. Cherry and D.E. Hooper, Amplifying Devices and Low-Pass Amplifier Design, John Wiley and Sons, New York, 1968.

5. The requirements for optimum performance in this respect fortunately appear to be to a large extent compatible with the requirements regarding the other quality aspects.

The design of bias circuitry is considered in chapter 6. It will be shown that this part of the design can be done in such a way that the signal-path performance of the amplifier is scarcely affected.

Finally, an outline of the design method is given in chapter 7. For examples of amplifiers designed according to the design procedure developed in this book reference is made to the literature.

This book is a revised and reviewed version of my Ph.D. thesis, which was published in June 1980 under the supervision of Prof.Dr.Ir. J. Davidse. The results of the work that he has encouraged me to carry out in the Laboratory of Electronics at the Delft University of Technology, The Netherlands, can be found here.

Writing such a thesis is perhaps even more of a burden to those in an author's environment than to himself. Without the aid of many others, it would not be possible to obtain a Ph.D. degree. As an acknowledgement of their support, this book is dedicated to all who contributed in some way.

Susan Masotty reviewed the text and corrected my numerous linguistic errors. Wim van Nimwegen drew the figures. Josette Verwaal and Hilda Verwest typed the manuscript. They thus contributed to the mere physical existence of this book, which happens to bear my name. This may, however, veil the fact that an author is no more than a person who is lucky enough to be able to write down the ideas he would never have had without a stimulating environment.

It is to this environment — which usually cannot be adequately indicated by names but includes the above — in all its abstraction that I am deeply indebted.

Delft, 1983 Ernst Nordholt

Contents

1 Basic Amplifier Configurations for Optimum Transfer of Information from Signal Sources to Loads

1.1 Introduction

One of the aspects of amplifier design most treated — in spite of its importance — like a stepchild is the adaptation of the amplifier input and output impedances to the signal source and the load. The obvious reason for neglect in this respect is that it is generally not sufficiently realised that amplifier design is concerned with the transfer of signal information from the signal source to the load, rather than with the amplification of voltage, current, or power.

The electrical quantities have, as a matter of fact, no other function than representing the signal information. Which of the electrical quantities can best serve as the information representative depends on the properties of the signal source and load. It will be pointed out in this chapter that the characters of the input and output impedances of an amplifier have to be selected on the grounds of the types of information representing quantities at input and output.

Once these selections are made, amplifier design can be continued by considering the transfer of electrical quantities. By speaking then, for example, of a voltage amplifier, it is meant that voltage is the information representing quantity at input and output. The relevant information transfer function is then indicated as a voltage gain.

After the discussion of this impedance-adaptation problem, we will formulate some criteria for optimum realisation of amplifiers, referring to noise performance, accuracy, linearity and efficiency. These criteria will serve as a guide in looking for the basic amplifier configurations that can provide the required transfer properties. The suitability of some feedback models will be discussed. The rather unusual, so-called asymptotic-gain model, will be selected for use in all further considerations.

Thereafter, we will present a classification of basic negative-feedback configurations. First, a rather theoretical approach is given, where the active amplifier part is considered as a nullor, while the feedback network is realised with ideal transformers and gyrators. A class of *non-energic* negative-feedback amplifiers

results, with up to maximally four overall feedback loops, forming the complete set of fundamentally different two-port amplifier types. More practical configurations with transformer feedback are treated next. Configurations with passive components (except transformers) in the feedback network are investigated in addition. These impedance feedback configurations are the most familiar amplifier types unfortunately not capable of realising all types of desired transfer functions. Therefore, amplifiers with active feedback are considered separately. These can provide the missing types of transfer functions. Two basically different types of active feedback can be distinguished. One of them will referred to here as indirect feedback.
These indirect-feedback amplifiers are especially suitable for realisation in integrated-circuit technology. As a practical restraint it is assumed throughout the whole chapter that source, load and amplifier have one common terminal ('ground').

1.2 General considerations

In figure 1.1 a general representation is given of a signal chain, where an amplifier is inserted between a signal source and a load. The input and output quantities of the source and load, respectively, are not necessarily electrical quantities. The three blocks in figure 1.1 can each be represented by either active or passive twoports. The parameters describing the transfer properties of these twoports will generally be functions of frequency, signal amplitude, temperature and other environmental circumstances. In order to realise an optimum information transfer from source to load, the transfer of each twoport should be determined by the parameter(s) that yield(s) *the best reproducing relation* between an input and an output quantity.

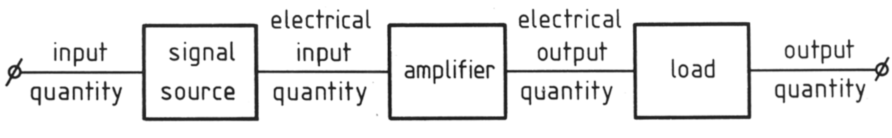


Figure 1.1 Amplifier inserted between signal source and load; source and load are formed by other electronic circuits or input and output transducers.

The best reproducing relation is determined in the case of transducers by their construction, their physical operation mechanism and — in the case of active transducers — by the way auxiliary power is supplied. The best reproducing input-output relation dictates the required character of the load impedance in the case of an input transducer (signal source) and of the source impedance for an output transducer (load).

An illustrative though simplified example of an input transducer is given below. A piezo-electric transducer produces a charge q_s linearly proportional to the pressure p, which is assumed to be the information representing primary signal. The best reproducing relation T of input and output quantities is therefore given by:

$$T = \frac{q_s}{p}$$

The output impedance of this transducer (the source impedance from the viewpoint of the amplifier) can be represented in the low-frequency region by a capacitance C_s. The open output voltage is given by:

$$u_{so} = \frac{q_s}{C_s}$$

This relation between voltage and charge is inaccurate, non-linear and temperature dependent, because of the properties of C_s. The open voltage does not therefore reproduce the primary information in an optimum way. The relation between short-circuit current and charge, however, is given by a linear differential equation:

$$i_{ss} = \frac{\mathrm{d}q}{\mathrm{d}t}$$

Therefore, a very low input impedance of the amplifier is required for optimum information transfer from transducer to amplifier. In other words, no voltage is allowed to arise across C_s for the transfer not to be affected by the less favourable properties of this source impedance.

Input transducers are preferably represented by Thévenin or Norton equivalent circuits. In that case, we have to deal with voltage and current sources only. The equivalent circuit should preferably be selected in accordance with the best reproducing relation of the transducer. In the case of the piezo-electric transducer, the Norton equivalent circuit is the obvious representation. It is shown in figure 1.2, where the current source is related to the charge.

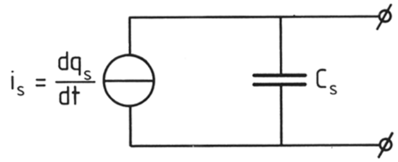


Figure 1.2 Preferable equivalent circuit for a piezo-electric transducer (in the low-frequency region).

If an accurate and linear relationship exists between open voltage and short-circuit current, as for example in *characteristic-impedance* systems, there is no preference for either one of the equivalent circuits. In that case *power* may be the information-representing quantity, and a possible criterion for optimum information transfer is that no power be reflected. To meet this criterion, an accurate and linear input impedance is needed.

Similar considerations can be given for loads. For instance, if the output quantity of the output transducer (load) has a linear and accurate relationship with the driving voltage, the amplifier should behave like a voltage source.

The transfer function of the amplifier which reproduces the information generated in the source as information supplied to the load will be labelled the *transmittance* of the amplifier. The various types of required transmittances are classified in table 1.1. Current, voltage and power are regarded as information-representing quantities. They can be derived quantities, as will be clear from the example of figure 1.2. To meet the requirements imposed by source and load, as far as the information transfer to and from the amplifier is concerned, it is necessary to give the input and output impedances their proper values, which may be either very large, very small, or accurate and linear. Various methods are available for realising the desired input and output impedances. They will be discussed in subsequent sections.

Table 1.1 Classification of different types of transmittances.

information-representing quantity produced by the source	information-representing quantity to be supplied to the load	transmittance
1. voltage	voltage	u_ℓ/u_s (voltage gain)
2. voltage	current	I_ℓ/U_s (transadmittance)
3. current	voltage	U_ℓ/I_s (transimpedance)
4. current	current	I_ℓ/I_s (current gain)
5. voltage	power	P_ℓ/U_s
6. power	voltage	u_ℓ/P_s
7. power	current	I_ℓ/P_s
8. current	power	P_ℓ/I_s
9. power	power	P_ℓ/P_s (power gain)

1.2.1 Preliminary criteria for the realisation of high-quality transmittances

Some general criteria on which the design of high-quality amplifiers has to be based are implicitly formulated below. Supporting material will be given in later chapters.

1. The insertion of series and shunt impedances in the signal path of an amplifier — causing a loss of available signal power — must be avoided. At the input, such a loss results in a relatively enlarged noise contribution of the active devices, while the resistors used add thermal noise to the signal. At the output, it results in a loss of efficiency and possibly increased non-linearity, because the active devices have to handle larger signal amplitudes.

On these grounds, methods for realising sufficiently high or low input or output impedances with the aid of passive series of shunt impedances will be called ‘brute force’ methods. A consistent application of the above formulation is one of the most important elements of this work.

2. To meet the requirements for optimum information transfer, the transmittance must be made independent of the properties of the active devices, and therefore the application of negative feedback is imperative. Power gain, produced by the active devices, has to be used as the expedient to realise both the desired *magnitude* of the transmittance and the desired *quality* of information transfer.

In this chapter we will deal with the basic negative-feedback amplifier configurations. These configurations can comprise one or more active devices in their *active parts*. When one active device is involved, we will call the configuration a *local-feedback* stage. When more stages are incorporated in the active part, it will be said that the configuration has *overall feedback*.
The loop gain in a multistage overall-feedback amplifier can be much larger than the loop gain in local-feedback stages. As a matter of fact, source and load impedances frequently have such characteristics that it will be impossible to realize the desired transmittance with a single active device. A cascade of local-feedback stages may then be considered, or alternatively, an overall feedback amplifier.
In this and the following chapters it will always be assumed that the feedback loop embraces as many stages as is consistent with optimum performance with respect to all quality aspects to be dealt with. High-frequency performance sets a limit to the useful magnitude of the loop gain and the number of stages in the active part. For solving certain amplifier problems it may be necessary to use two or more cascaded overall-feedback amplifiers. In that case, the designer has the freedom to choose the characters of the first-amplifier output impedance and the second-amplifier input impedance. Cherry and Hooper [1] have shown that, in order to avoid undesired interaction, an impedance mismatch between the amplifiers is favourable. We will not deal with this aspect in this work any further.

For the design of negative-feedback amplifiers, it is desirable to have at one's disposal a feedback model which closely conforms to the above view. In the next section we will briefly discuss some methods for the analysis of feedback amplifiers and the suitability of these methods for design purposes.

1.3 Models for negative-feedback amplifiers

1.3.1 Black's feedback model

Though not necessary, it is frequently considered appropriate to make the presence of negative feedback in an amplifier explicit. For design purposes, one can think about negative feedback as a means of exchanging available power gain for quality of information transfer. This view comes somewhat to the fore in the *elementary feedback model* of Black's feedback patent [2] given in figure 1.3. This model is commonly used as a design aid. The transfer function is given by:

$$A_t = \frac{E_\ell}{E_s} = \frac{A}{1 - Ak} \qquad (1.1)$$

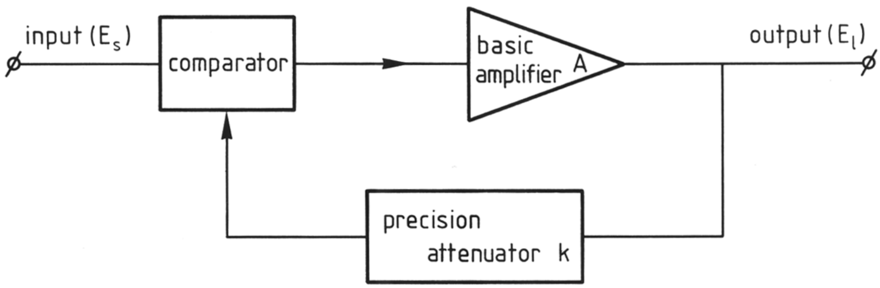


Figure 1.3 Elementary feedback model.

Because the model is a block diagram based on transfer functions which are assumed to be unilateral and not affecting each other, it is hardly applicable to practical amplifier configurations. The calculation of the transmittance of a real amplifier is therefore usually performed with the aid of matrix-parameter representations of the amplifier and feedback network [3], [4]. As a result, an identification problem arises when the expression for A_t found in this way is manipulated into the form of relation (1.1).

As a matter of fact, this method is suitable for analysis rather than for straightforward design. It is applicable to a given configuration but gives no explicit indications for the selection of a configuration for a certain purpose. Besides, the use of this model bears the risk that one might try and realise the transfer functions A and k in such a way that they are indeed unilateral and not affecting each other. As a design philosophy for general purpose operational amplifiers, this may be a defensible approach, but as will become apparent from the subsequent chapters, it does not lead to optimum special-purpose amplifier configurations.

1.3.2 Anti causal analysis

Recently, Waldhauer [5] has proposed the so-called *anti causal analysis* of feedback amplifiers. This analysis is made with the aid of the *transmission* matrix, where the input quantities are expressed as functions of the output quantities. The reciprocal value of the amplifier transfer function is found by this method to be the sum of a number of terms. The most important term describes the design goal and is determined by the components of the feedback network. The other terms express the non-ideality of the active part. Though the presence of feedback is not made explicit, the concept of feedback is used to find the basic configuration for a certain application. This method is, like the previous one, suitable for analysis rather than for design.

1.3.3 Superposition model

An equally formal approach of feedback-amplifier analysis leads to a model that can be indicated as the *superposition model*, because it is derived by using the superposition principle. This model will be derived here, because it forms the basis for the model that will appear to be most suitable for our design purposes.
Each voltage and current in the amplifier can be written as a linear combination of the quantity produced by the signal source and the current or voltage of an arbitrarily selected controlled source in the amplifier. For setting up the network equations, the controlled source is initially regarded as independent. If we indicate the quantities regardless of their dimensions by the symbol E, the network equations for the load and input quantities can be written as:

$$E_\ell = \rho E_s + \nu E_c \qquad (1.2)$$

$$E_i = \zeta E_s + \beta E_c \qquad (1.3)$$

The indices ℓ, i, s and c refer to load, input, signal source and controlled source, respectively. Besides, there is a relation between E_c and E_i. that can be written as:

$$E_c = AE_i \qquad (1.4)$$

A non-zero value of the transfer function β indicates the existence of feedback. The transfer functions ρ, ν, ζ, A and β can have dimensions [Ω], [1/Ω] or are dimensionless. The transfer function A is called the *reference variable*.
The transfer function A_t of the amplifier is found from (1.2), (1.3) and (1.4):

$$A_t = \frac{E_\ell}{E_s} = \rho + \nu\zeta\frac{A}{1 - A\beta} \tag{1.5}$$

The product $A\beta$ is called the *loop gain* with respect to the reference variable A, while $1 - A\beta$ is the well-known *return difference,* with respect to the reference variable A, as defined by Bode [6].
An example may elucidate how A_t is calculated. Figure 1.4 shows an arbitrarily chosen, very simple configuration, where the amplifier is represented by a current source controlled by the input voltage. The transconductance g is selected for obvious reasons as the reference variable. Calculating now the factors

$$\rho = (U_\ell/I_s)_{I_c=0}, \quad \nu = (U_\ell/I_c)_{I_s=0}, \quad \zeta = (U_i/I_s)_{I_c=0} \quad \text{and} \quad \beta = (U_i/I_c)_{I_s=0}$$

we find for the transfer function A_t:

$$A_t = \frac{U_\ell}{I_s} = \frac{Z_iZ_\ell - gZ_fZ_iZ_\ell}{Z_\ell + Z_f + Z_i + gZ_iZ_\ell}$$

If the loop gain $A\beta = -gZ_iZ_\ell/(Z_\ell + Z_f + Z_i)$ is very large, A_t can be approximated by:

$$A_t = -Z_f$$

The transfer function stabilised by the negative feedback in this case is a transimpedance.

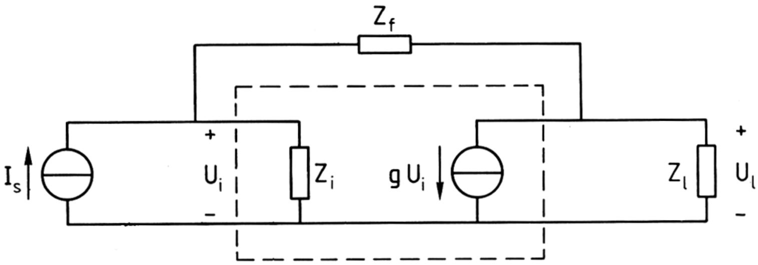


Figure 1.4 Example of a feedback configuration.

Although this superposition model offers the possibility to exactly calculate the transmittance in a relatively simple way, in contrast with other models or calculation methods, once again it is only suitable for analysis. Some qualitative design indications can be derived from the transfer functions ρ, ζ and ν. The direct transfer from source to load is described by ρ, and ζ and ν are measures for the signal loss at input and output, respectively. The model does not, however, provide any explicit indications for the realisation of high-quality transmittances. We have paid so much attention to it, because the so-called *asymptotic-gain model* can easily be derived from it and because the latter model does not have the disadvantages mentioned earlier.

1.3.4 Asymptotic-gain model

(i) The transfer functions

If in (5) the loop gain becomes infinite ($A\beta \rightarrow \infty$), the transfer function is given by:

$$\lim_{A\beta \rightarrow \infty} A_t = A_{t\infty} = \rho - \frac{\nu\zeta}{\beta} \tag{1.6}$$

where A_t will be referred to as the *asymptotic gain*. Substitution of (1.6) in (1.5) yields:

$$A_t = \frac{E_\ell}{E_s} = \frac{\rho}{1 - A\beta} - A_{t\infty}\frac{A\beta}{1 - A\beta} \tag{1.7}$$

In all practical amplifier designs, the first term in (1.7) will be much smaller than the second, and (1.7) can be written as:

$$A_t = \frac{E_\ell}{E_s} = A_{t\infty}\frac{A\beta}{1 - A\beta} \tag{1.7}$$

The design of the transfer function of a negative-feedback amplifier is thus reduced to two successive steps. The first step is the determination of $A_{t\infty}$, which can be considered as the design goal, and the second step is the realisation of an adequate loop transfer function $A\beta$.

Unless the input and/or output impedances need to be accurate, it is not necessary to calculate their values. The amplifier performance is adequately described by expression (1.8). Note that the transmittance is defined as the transfer function:

$$A_t = \frac{\text{load quantity}}{\text{source quantity}} \qquad \text{and not} \qquad \frac{\text{output quantity}}{\text{input quantity}}$$

To ascertain the meaning of $A_{t\infty}$ we consider the expression for E_i, which follows from (1.3) and (1.4):

$$E_i = \frac{\zeta}{1 - A\beta} E_s \tag{1.9}$$

Obviously, E_i approaches zero when $|A\beta/\zeta|$ becomes infinite. If for E_i the current or voltage at the input of the first amplifier stage is selected, E_i will be constrained to be zero when the loop gain becomes infinite. If, for example, E_i is the voltage between two nodes connected by one branch (as in the example of figure 1.3) then the current in that branch will also approach zero when $A\beta$ becomes infinite. The constraints correspond to the *nullor* [7] constraints in network theory.

(ii) Impedance calculation

The superposition principle can also be used to calculate input and output impedances of the amplifier. Only in the case where these impedances are intended to become accurate and linear by the feedback action will the calculation be useful. We will follow here the treatment as given by Boon [8].

Instead of E_ℓ, which is the load quantity in the case of the gain calculation, we will in the network equations now write E'_s in order to emphasise that the driving quantity E_s and its response E'_s are related to one and the same port. The relevant network equations can then be written as:

$$E'_s = \rho E_s + \nu E_c \tag{1.10}$$

$$E_i = \zeta E_s + \beta E_c \tag{1.11}$$

With the additional relation

$$E_c = AE_i \tag{1.12}$$

and when the relation to be found between E_s and E'_s is an impedance Z_t according to

$$E'_s = Z_t E_s \tag{1.13}$$

we find the following expression for the impedance function:

$$Z_t = \rho \frac{1 - A(\beta - \frac{\nu\zeta}{\rho})}{1 - A\beta} \tag{1.14}$$

Defining now:

$$\beta_o = \left(\frac{E_i}{E_c}\right)_{E_s=0} \tag{1.15}$$

$$\beta_{sc} = \left(\frac{E_i}{E_c}\right)_{E'_s=0} \tag{1.16}$$

the impedance function (1.14) can he rewritten as

$$Z_t = \rho \frac{1 - A\beta_{sc}}{1 - A\beta_o} \tag{1.17}$$

which is Blackman's [9] formula.

Two loop gains $A\beta_{sc}$ and $A\beta_0$ may play a role in the determination of an input or output impedance. In amplifiers with a single feedback loop (to be dealt with in the next section) one of both loop gains is zero while the other is large so that the impedances tend either to zero or infinity. We have seen before that there is no reason to calculate their values because their effects on the amplifier transfer are included in the expression (1.8) for A_t.

In order to realize an accurate impedance, both loop gains have to be large which, as we shall see in the next section, requires at least two feedback loops.

In the ideal case of infinite loop gains (or, in other words, when the active part is assumed to have nullor properties), we find that the impedance obtains a value

$$Z_{t\infty} = \lim_{\substack{A\beta_o \to \infty \\ A\beta_{\sigma\chi} \to \infty}} Z_t = \rho \frac{\beta_{sc}}{\beta_o} \tag{1.18}$$

Expressions (1.17) and (1.18) can be combined to set up an expression in which the impedance is the product of its asymptotic value $Z_{t\infty}$ and of two factors accounting for the non-ideality of the active part, the same as in the asymptotic-gain model. The same loop gains occur, of course, as in Blackman's formula (1.17):

$$Z_t = Z_{t\infty} \frac{-A\beta_o}{1 - A\beta_0} \cdot \frac{1 - A\beta_{sc}}{-A\beta_{sc}} \tag{1.19}$$

This expression can be simplified for those cases where the impedance ρ — the impedance in the case where $A = 0$ — strongly differs from the desired impedance $Z_{t\infty}$. This will usually be true for amplifiers with two feedback loops. In the case where $|\rho| \ll |Z_{t\infty}|$, we find:

$$Z_t = Z_{t\infty} \frac{-A\beta_{sc}}{1 - A\beta_{sc}} \tag{1.20}$$

In the case where $|\rho| \gg |Z_{t\infty}|$, we find:

$$Z_t = Z_{t\infty} \frac{1 - A\beta_{sc}}{-A\beta_{sc}}$$

which can alternatively and for the sake of more uniformity be written as:

$$Y_t = Y_{t\infty} \frac{-A\beta_{sc}}{1 - A\beta_{sc}} \tag{1.21}$$

Expressions (1.20) and (1.21) have the same form as the expression for the gain. However, different loop gains determine the deviations from the intended ideal values of gain and impedance (or admittance). For design purposes it is useful to establish the relation between the two loop gains, which will be done below.

(iii) Relations between transfer and impedance functions

The loop gain $A\beta$ of the amplifier transfer function has to be calculated with the amplifier ports terminated in the source and load impedances. When we indicate the external impedance as Z, we can, according to Blackman [9], write:

$$A\beta = \frac{\rho}{\rho + Z} A\beta_{sc} + \frac{Z}{\rho + Z} A\beta_o \tag{1.22}$$

Using expression (1.18) we can find the desired relations

$$A\beta = \frac{1}{\rho + Z} (Z_{t\infty} + Z) A\beta_o \tag{1.23}$$

$$A\beta = \frac{\rho}{\rho + Z} \left(1 + \frac{Z}{Z_{t\infty}}\right) A\beta_{sc} \tag{1.24}$$

Equation (1.23) is useful in the case where $|\rho| \ll |Z_{t\infty}|$ while equation (1.24) is more suitable for situations where $|\rho| \gg |Z_{t\infty}|$.

These equations can be simplified when Z and $Z_{t\infty}$ have the same order of magnitude, which is the case with characteristic impedance terminations. We then find:

$$A\beta = \left(1 + \frac{Z_{t\infty}}{Z}\right) A\beta_o \tag{1.25}$$

$$A\beta = \left(1 + \frac{Z}{Z_{t\infty}}\right) A\beta_{sc} \tag{1.26}$$

These equations can be used in many situations for a comparison between the impedance and transfer behaviour, in particular of amplifiers with two feedback loops.

Summarising, we can conclude that it is possible to calculate input and output impedances in a way similar to that for the amplifier gain. Clearly, we can determine the ideal transfer function $A_{t\infty}$ together with the ideal impedances $Z_{t\infty}$ as if the active part were a perfect nullor. *The first design step, which involves the determination of the basic amplifier configuration* is thus considerably simplified.

In the subsequent part of this chapter we will make an inventory of basic amplifier configurations, assuming the active part to have nullor properties. Design aspects regarding the loop transfer function(s) $A\beta$ will be put off until later chapters.

1.4 The realisation of transmittances with passive feedback networks

1.4.1 Introduction

The problem of realising the various transmittances as given in table 1.1 might be approached as a mere network-theoretical structural synthesis problem. Such an approach, however, easily loses sight of the practical electronic realisation. We will therefore keep in mind that the active part of a negative-feedback amplifier cannot really be a nullor, but that it produces noise and distortion, that it dissipates power, that it has no perfectly floating ports and that its parameters are frequency dependent. Nevertheless, we will, as far as its transfer properties are concerned, attribute nullor properties to the active part in this and the following sections of this chapter, in order to find the basic amplifier configurations. These configurations will thus be described with the asymptotic gain as defined in section 1.3. In the subsequent chapters we will take full account of the imperfections in the active part.

The elements of the feedback network will be assumed to be ideal and for the time being we will admit all passive network elements, namely the resistance, capacitance, inductance, transformer and gyrator. Though the practical significance of some configurations may be small, this approach is considered to give the best fundamental insight into the electronic realisation problem.

In the next section we will present realisations of all types of transmittances, following a mainly network-theoretical approach. In later sections we will turn our attention to practical realisations, making use of the insight we can gain from general theory. First, we will give some definitions and descriptions that will be useful for our classification purposes.

(i) Some definitions

In this work we will frequently use the *transmission parameters* to describe the transfer properties of a twoport. The twoport equations express the input quantities as functions of the output quantities in the following way:

$$U_i = AU_o + BI_o \qquad I_i = CU_o + DI_o \quad \text{or} \quad \begin{pmatrix} U_i \\ I_i \end{pmatrix} = \begin{pmatrix} A & B \\ C & D \end{pmatrix} \begin{pmatrix} U_o \\ I_o \end{pmatrix}$$

Signs are attributed to the quantities as indicated in figure 1.5.

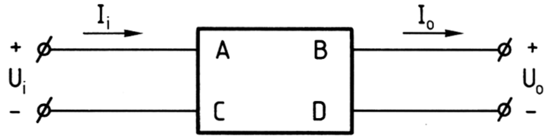


Figure 1.5 Sign conventions for twoports described by their transmission parameters.

Somewhat more familiar to the electronic designer are the reciprocal values of the transmission parameters, which will be referred to as the *transfer parameters*. They are defined as follows:

Voltage-gain factor $\mu = \frac{1}{A} = \left(\frac{U_o}{U_i}\right)_{I_o=0}$

Transadmittance $\mu = \frac{1}{B} = \left(\frac{I_o}{U_i}\right)_{U_o=0}$

Transimpedance $\mu = \frac{1}{C} = \left(\frac{U_o}{I_i}\right)_{I_o=0}$

Current-gain factor $\mu = \frac{1}{D} = \left(\frac{I_o}{I_i}\right)_{U_o=0}$

These definitions implicitly show how to measure the transfer parameters and how to find, indirectly, the transmission parameters.

The active part of the amplifier will — in accordance with the asymptotic gain model — be modelled as a nullor. All transmission parameters of a nullor are zero; all transfer parameters are infinite.

(ii) Description of feedback techniques

In order to find the various basic amplifier configurations with fundamentally different transfer parameters, it is necessary to accurately describe the techniques of applying negative feedback. We will do this first.

The application of negative feedback requires that the output quantity that has to be supplied to the load be sensed by the input port of the feedback network.

– If, for obtaining optimum signal transfer, the load has to be driven from a voltage source, then the voltage across the load must be sensed. This type of feedback is generally called *output shunt feedback*.

- Current sensing, commonly designated as *output series feedback,* is used when the amplifier has to behave as a current source.
- A combination of series and shunt feedback at the output is required when the output power has to be sensed, for example, in order to accomplish characteristic impedance matching.

The sensed output quantity has to be converted accurately in the feedback network into a quantity corresponding to the source quantity that represents the signal information. A comparison is made by summing or subtracting these two quantities at the input of the active part of the amplifier.

- When these quantities are currents, the type of feedback is called *input shunt feedback*.
- In the case of voltages the term *input series feedback* is used.
- In order to realize, for example, characteristic impedance matching a combination of series and shunt feedback at the input is needed.

1.4.2 Classification of non-energic feedback amplifiers

The general criteria formulated in section 1.2 imply the desirability of such a realisation of the feedback action, so that no shunt or series impedances in the signal path are needed. In practice, this cannot be accomplished perfectly, but theoretically, the required feedback networks are available in the form of the ideal transformer and the ideal gyrator. A transformer provides the possibility of converting a voltage into a voltage and a current into a current. The gyrator converts a voltage into a current and vice versa. The gyrator and the transformer are both *non-energic* [10] network elements. No instantaneous power absorption occurs. Noise performance and power efficiency are not degraded when only these ideal elements are used in the feedback network. This was already pointed out by Norton [11] for some transformer-feedback circuits.

Figure 1.6 gives the symbols of the non-energic network elements, together with their transmission matrices.

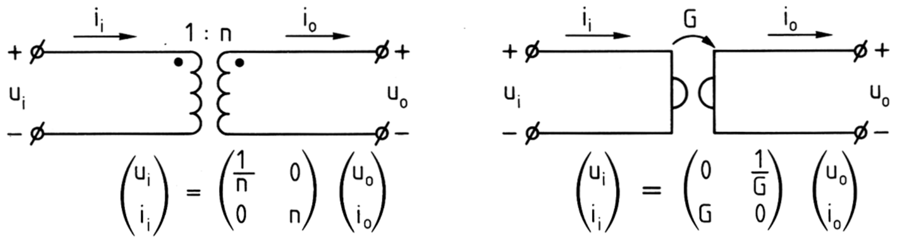


Figure 1.6 *Network symbols and transmission matrices of the ideal transformer and gyrator.*

To restrict the number of possible basic-amplifier realisations, the input and output ports of the active part will be given one common terminal, which is also in common with the source and load. Because the source, load and active part have one-side grounded ports, the ports of the feedback elements must be able to float in order to accomplish current sensing and voltage comparison (series feedback at output and input, respectively).

Four feedback loops can be applied to the active part of an amplifier, as indicated in figure 1.7. The feedback elements accomplish the following conversions:

Transformer n_1	Voltage-to-voltage,
Gyrator G_1	Current-to-voltage,
Gyrator G_2	Voltage-to-current,
Transformer n_2	Current-to-current.

We will assume that loop gain is infinite. Consequently, the active part has nullor properties. Its transfer parameters are assumed to have negative signs.

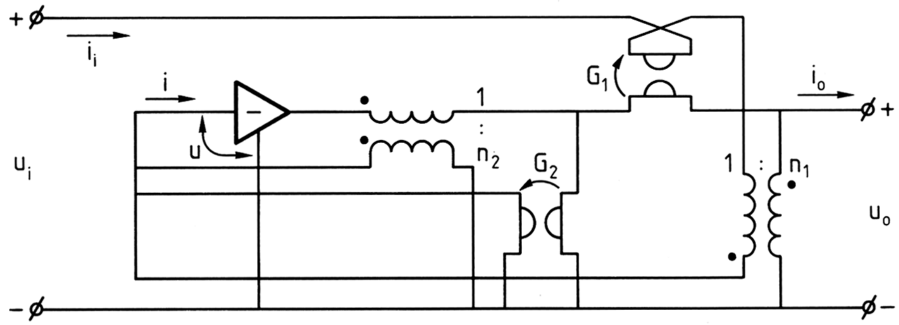


Figure 1.7 *Amplifier with the maximum number of feedback loops. All transfer parameters are determined by the feedback-network transfer function.*

The order of the various sensings is chosen arbitrarily. The exact values of the transfer parameters for this specific order are:

$$\mu = -n_1 \left\{ \frac{1 + G_2/G_1 + 1/n_1 n_2}{1 + 1/n_1 n_2} \right\}$$

$$\zeta = -\frac{1}{G_2} \{1 + G_2/G_1 + 1/n_1 n_2\}$$

$$\gamma = -G_1 \left\{ \frac{1 + G_2/G_1 + 1/n_1 n_2}{1 + G_2/G_1} \right\}$$

$$\alpha = -n_2 \{1 + G_2/G_1 + 1/n_1 n_2\}$$

If all these transfer parameters are large, they are approximately independently determined by the feedback elements, because in that case the direct transfer of current and voltage to the output is small with respect to the transfer via the active part.

By eliminating one or more gyrators or transformers from figure 1.7, less complicated feedback configurations having three loops or less are obtained. Sixteen possible configurations result. Table 1.2 list their *approximate* transfer parameters, assuming that $n_1 n_2 \gg 1$ and $G_1/G_2 \gg 1$.

When these approximations are not allowed, the *exact* expressions of the transfer parameters can easily be found from the above exact expressions for the four-loop amplifier.

Table 1.2. Transfer parameters of the nullor and non-energic feedback configurations with up to four feedback loops (approximate values are given for configurations 6 – 16).

	$-\frac{1}{A} = -\mu$	$-\frac{1}{B} = \gamma$	$-\frac{1}{C} = -\zeta$	$-\frac{1}{D} = -\alpha$	
1	∞	∞	∞	∞	nullor
2	n_1	∞	∞	∞	one loop
3	∞	G_1	∞	∞	
4	∞	∞	$1/G_2$	∞	
5	∞	∞	∞	n_2	
6	n_1	G_1	∞	∞	two loops
7	n_1	∞	$1/G_2$	∞	
8	n_1	∞	∞	n_2	
9	∞	G_1	$1/G_2$	∞	
10	∞	G_1	∞	n_2	
11	∞	∞	$1/G_2$	n_2	
12	∞	G_1	$1/G_2$	n_2	three loops
13	n_1	∞	$1/G_2$	n_2	
14	n_1	G_1	∞	n_2	
15	n_1	G_1	$1/G_2$	∞	
16	n_1	G_1	$1/G_2$	n_2	four loops

Note that each feedback loop essentially fixes the value of one transfer parameter. The properties of the amplifiers will be studied in some detail by inserting them between a signal source and a load as shown in figures 1.8a and 1.8b, where the signal sources are given as voltage and current sources, respectively.

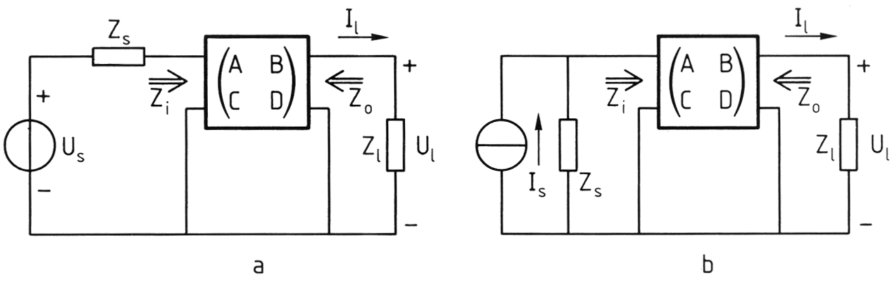


Figure 1.8 Amplifiers inserted between signal sources and loads.

The voltage transfer function of figure 1.8a is given by:

$$\frac{U_\ell}{U_s} = \frac{Z_\ell}{AZ_\ell + B + CZ_s + DZ_s} = A_u \tag{1.27}$$

where A, B, C and D are the transmission parameters of the amplifier. The other transfer functions are related to A_u as:

$$\backslash f(I_\ell,U_s) = \frac{A_u}{Z_\ell} \quad \text{(figure 1.8a)}$$

$$\backslash f(U_\ell,I_s) = A_u Z_s \quad \text{(figure 1.8b)}$$

$$\backslash f(I_\ell,I_s) = A_u \frac{Z_s}{Z_\ell} \quad \text{(figure 1.8b)}$$

The input impedance Z_i and the output impedance Z_o are given respectively by:

$$Z_i = \frac{AZ_\ell + B}{CZ_\ell + D} \quad \text{and} \quad Z_o = \frac{B + DZ_s}{A + CZ_s}$$

The specific properties with respect to the information-transfer aspects of the various configurations will be discussed in the following.

(i) The nullor

Without external circuitry, the nullor is obviously not suitable for the transfer of signal information.

(ii) Single-loop configurations (4)

The four single-loop amplifiers have either zero or infinite input and output impedances and a single well-determined transmission parameter. They can be used when the signal information is represented by a voltage or a current and when the load has to be driven from a voltage or a current source. The source and load impedances have no influence on the transfer functions. Signal-to-noise ratio (SNR) and power efficiency (PE) are exactly equal to those of the active part of the amplifier. This may be elucidated by an example.

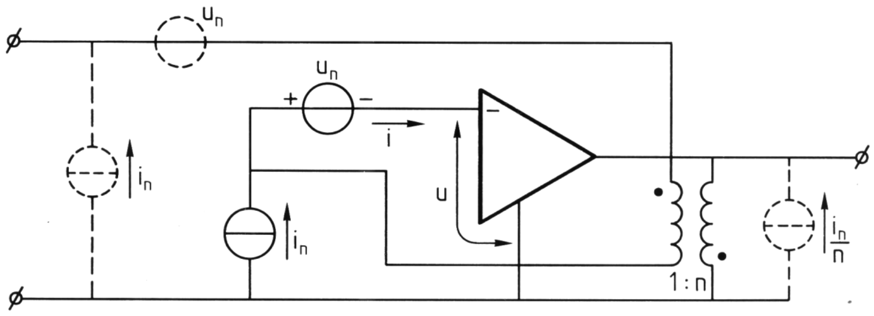


Figure 1.9 Voltage amplifier with input noise sources of the active part and transformed noise sources indicated by dashes.

Figure 1.9 shows a voltage-amplifier configuration, with a zero output impedance, an infinite input impedance and a transfer parameter $\mu = -1/n$. All other transfer parameters are infinite. The equivalent input noise sources u_n and i_n of the active part are given in the figure. The result of a transformation of these noise sources is given in the same figure. The transformation techniques used are explained in section 3.2.

The source i_n/n does not contribute to the equivalent input noise sources, because it is transformed into input sources by dividing it by the transfer parameters α and γ, which are both infinite. The equivalent input noise sources are consequently equal to those of the active part. Furthermore, there is no power loss at the output, because the current in the transformer equals zero. Similar considerations can be given for the other single-loop configurations.

(iii) Dual-loop configurations (6)

The combination of shunt and series feedback either at the input or at the output or at both input and output leads to the possibility of realising accurate power transfer which is needed for characteristic impedance matching.

The input impedance may or may not depend on the load impedance. If it does, the amplifier is not *unilateral,* and as a consequence, accurate matching at the input port is possible only if the load impedance is accurately known. If it does not, the matching at the input port is independent of the situation at the other port, where the driving point impedance is either zero or infinite. The output impedance exhibits a similar behaviour, in relation to the source impedance.

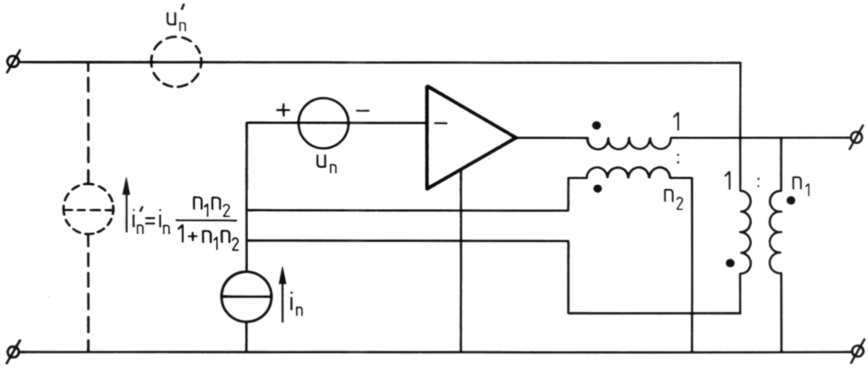


Figure 1.10 Transformation of the noise sources of the active part to the input of a dual-loop amplifier.

In some cases, the feedback networks cause some improvement of the SNR compared to that of the active part. An example is given in figure 1.10, where the equivalent input noise sources of the active part are transformed into input noise sources of the amplifier. The noise current source i'_n is somewhat smaller than i_n, while the sources u'_n and u_n are- equal. The resulting SNR improvement (too small to be useful) is a consequence of the increase in available output power by the direct transfer, via the feedback network, from source to load. This relative share of the directly transferred source power in the available output power gets smaller as the available power gain gets larger. The impedance and transfer properties of all possible (6) dual-loop feedback amplifier types are summarised in table 1.3. The configuration numbers refer to table 1.2. Configurations 8 and 9 can be used for characteristic impedance matching at input and output ($Z_i = Z_s$, $Z_o = Z_\ell$), because $Z_iZ_o = Z_\ell Z_s$ and $Z_iZ_\ell = Z_oZ_s$, respectively. For equal impedance levels at input and output, A and D must be equal in configuration 8, while in configuration 9 the condition that $B/C = Z_sZ_\ell$ must be met. Matching errors at one port are reflected at the other port. As a consequence, these configurations are not ideal for matching purposes.

(iv) Three-loop configurations (4)

Though it is possible to realize accurate input and output impedances with these configurations when load and source impedances are accurately known, there is no possibility for characteristic matching at both sides. The interdependence of Z_i and Z_ℓ and of Z_o and Z_s cannot be avoided. The configurations are therefore not attractive for any application.

Table 1.3 Transfer properties of dual-loop feedback amplifiers.

					Z_i	Z_o	transfer function
6	A	B			∞	$\frac{B}{A}$	$\frac{U_\ell}{U_s} = \frac{Z_\ell}{AZ_\ell + B}$
7	A		C		$\frac{A}{C}$	0	$\frac{U_\ell}{I_s} = \frac{Z_s}{A + CZ_s}$
8	A			D	$\frac{AZ_\ell}{D}$	$\frac{DZ_s}{A}$	$\frac{U_\ell}{U_s} = \frac{Z_\ell}{AZ_\ell + DZ_s}$
9		B	C		$\frac{B}{CZ_\ell}$	$\frac{B}{CZ_s}$	$\frac{U_\ell}{U_s} = \frac{Z_\ell}{B + CZ_s/Z}$
10		B		D	$\frac{B}{D}$	∞	$\frac{I_\ell}{U_s} = \frac{1}{B + DZ_s}$
11			C	D	0	$\frac{D}{C}$	$\frac{U_\ell}{I_s} = \frac{Z_\ell}{CZ_\ell + D}$

(v) Four-loop configuration (1)

All parameters are determined accurately in the case of four feedback loops. The parameters can be designed so that input and output impedances do not depend on source and load impedances. The condition for this independence is:

$$AD = BC, \quad \text{or} \quad \zeta\gamma = \mu\alpha$$

In that case:

$$Z_i = A/C = B/D \quad \text{and} \quad Z_o = B/A = D/C$$

Impedance matchings at input and output are not interdependent. With characteristic matching at both sides, the voltage transfer function is given by:

$$\frac{U_\ell}{U_s} = \frac{1}{4A}$$

(vi) Concluding remarks

The configurations which were presented in this section are paradigms of negative-feedback amplifiers, where sensing of the output quantities and comparison of the input quantities are accomplished in an ideal, non-energic way. Moreover, each feedback loop virtually independently determines one transfer parameter provided that the direct power transfer from source to load is relatively small. The theoretical significance of this classification is great because the configurations form a complete set of ideal amplifier types. Their practical significance is slight because the ideal gyrator cannot be approximated well enough.

As soon as impedances are used in the feedback networks, the ideal situation is lost. For example, current sensing will cause a voltage drop and voltage sensing will cause extra current flow. Noise may be generated in the feedback impedances and SNR and PE may be deteriorated. The class of non-energic feedback amplifiers admits realisations of transfer functions with either positive or negative signs, thanks to the floating ports of the feedback networks.

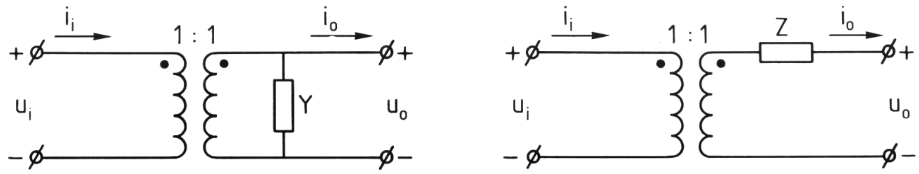
Practical amplifiers must necessarily have feedback networks with floating ports too, in order to obtain transfer properties similar to the configurations discussed. If active feedback networks are not considered, these practical realisations must use transformers for obtaining inverting as well as non inverting alternatives for the gyrator feedback. A small number of transfer function types can be realised without transformers. The class of transformer feedback configurations is presented in the next section.

1.4.3 Classification of amplifier configurations with transformer feedback

A physical realisation of the transformer can approximate the ideal transformer to a large extent, provided that the impedance and power levels at which it is used are low, that its physical dimensions are small (low stray capacitance) and that frequencies are not too low. The transformer (e.g. wound on a toroid core) can then be modelled over a number of frequency decades as an ideal transformer with floating ports.

The gyrator is much more difficult to realize in a useful physical form. Electronic realisations may approximate the ideal in a few low-frequency decades, but only as far as non-energicness is concerned. Their noise production and non-linearity — inherent when active devices and resistances that fix the gyration constants are used — render them useless as feedback elements in practical wide-band amplifiers.

The gyrator function in the feedback network can be performed, however, by impedances in combination with transformers. In some cases the transformers can be omitted. Figure 1.11 shows the alternatives for the gyrator with the corresponding transmission matrices. For the sake of simplicity the transformers will have unity turns ratios. The active part of the amplifier will have one-side-grounded ports again and will invert the input quantities (μ, γ, ζ and α all have values $-\infty$). In some special cases we will forego this restriction.



$$\begin{pmatrix} u_i \\ i_i \end{pmatrix} = \begin{pmatrix} 1 & 0 \\ Y & 1 \end{pmatrix} \begin{pmatrix} u_o \\ i_o \end{pmatrix} \qquad \begin{pmatrix} u_i \\ i_i \end{pmatrix} = \begin{pmatrix} 1 & Z \\ 0 & 1 \end{pmatrix} \begin{pmatrix} u_o \\ i_o \end{pmatrix}$$

Figure 1.11 Transformer-immittance combinations to be used as practical alternatives for the gyrator.

The one-sided grounding of the ports of the active part has a practical significance now. Perfectly floating ports are hard to realize. An infinite value of the common-mode rejection ratio is needed at the input, and perfectly equal currents have to flow in the output leads. Parasitic effects result if these conditions are not met. Input and output impedances are shunted in that case by parasitic impedances, and distortion may be enhanced. We will take these imperfections into account in chapter 4. Figure 1.12 depicts the configuration with four feedback loops. These are not completely independent now. Current sensing, for example, influences the voltage sensing and vice versa.

The following equations hold if nullor properties are attributed again to the active part:

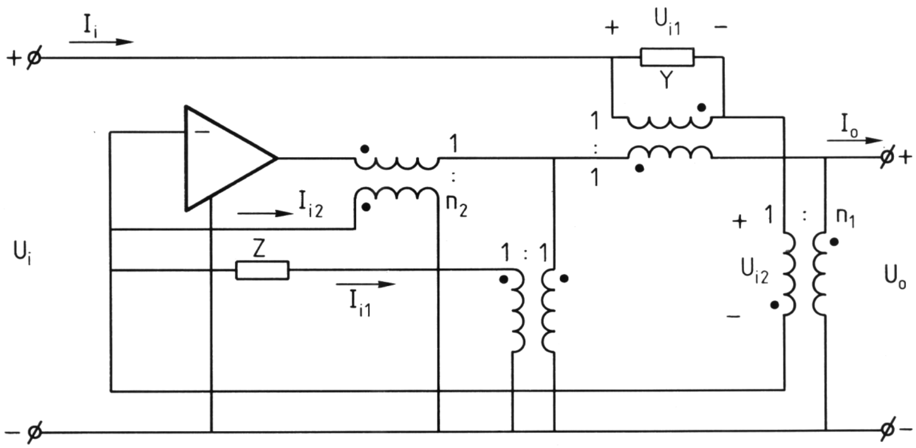


Figure 1.12 General feedback configuration with four feedback loops implemented with transformers and immittances.

$$U_i = U_{i1} + U_{i2}, \quad I_i = I_{i1} + I_{i2}$$

$$-I_{i1}\left(\frac{n_1 - 1}{n_1}\right) + I_{i2}\left(\frac{n_1 n_2 + 1}{n_1}\right) = -I_o$$

$$U_o - U_{i1} = -I_{i1}Z, \quad U_o = -n_1 U_{i2}, \quad U_{i1} = \frac{I_i}{Y}\left(\frac{n_1 - 1}{n_1}\right) - \frac{I_o}{Y}$$

All parameters of all 16 possible configurations can be determined from these equations. We will discuss their properties briefly below.

(i) The active part of the amplifier

The active part of the amplifier will generally not be suitable for information transfer unless additional circuitry is used. It produces power gain in order to realize the desired magnitude of the transfer function as well as to obtain a high quality of information transfer when used in combination with one or more negative-feedback loops.

(ii) Single-loop configurations (figure 1.14)

The four single-loop amplifiers, derived directly from figure 1.12 by making three out of the four quantities Y, Z, n_1 and n_2 infinite, have transfer parameters that are respectively given by:

Voltage amplifier	$\mu = -n_1$,
Transadmittance amplifier	$\gamma = -Y$,
Transimpedance amplifier	$\zeta = -Z$,
Current amplifier	$\alpha = -n_2$.

The transformer can be omitted in the transimpedance amplifier, provided that the active part is inverting. The μ and α realisations need transformers exclusively and are therefore essentially non-energic. Input and output impedances of the configurations are either zero or infinite.

Non-inverting unity voltage-gain and unity current-gain amplifiers can be realised also without the aid of transformers provided that the amplifier has floating ports. The type of feedback is non-energic, but they have the disadvantages of parasitic effects, as mentioned before. Figure 1.13 shows the symbol adopted for an active part with floating ports. Figure 1.14 shows the single-loop configurations with their more or loss currently used names.

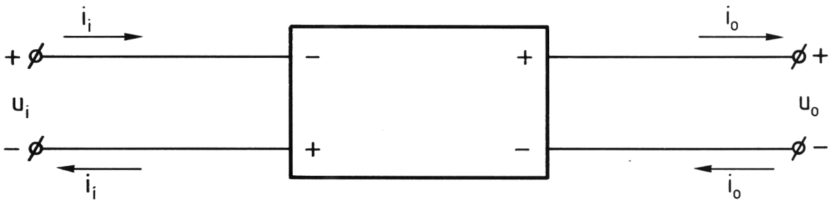


Figure 1.13 *Symbol for ideally floating active part.*

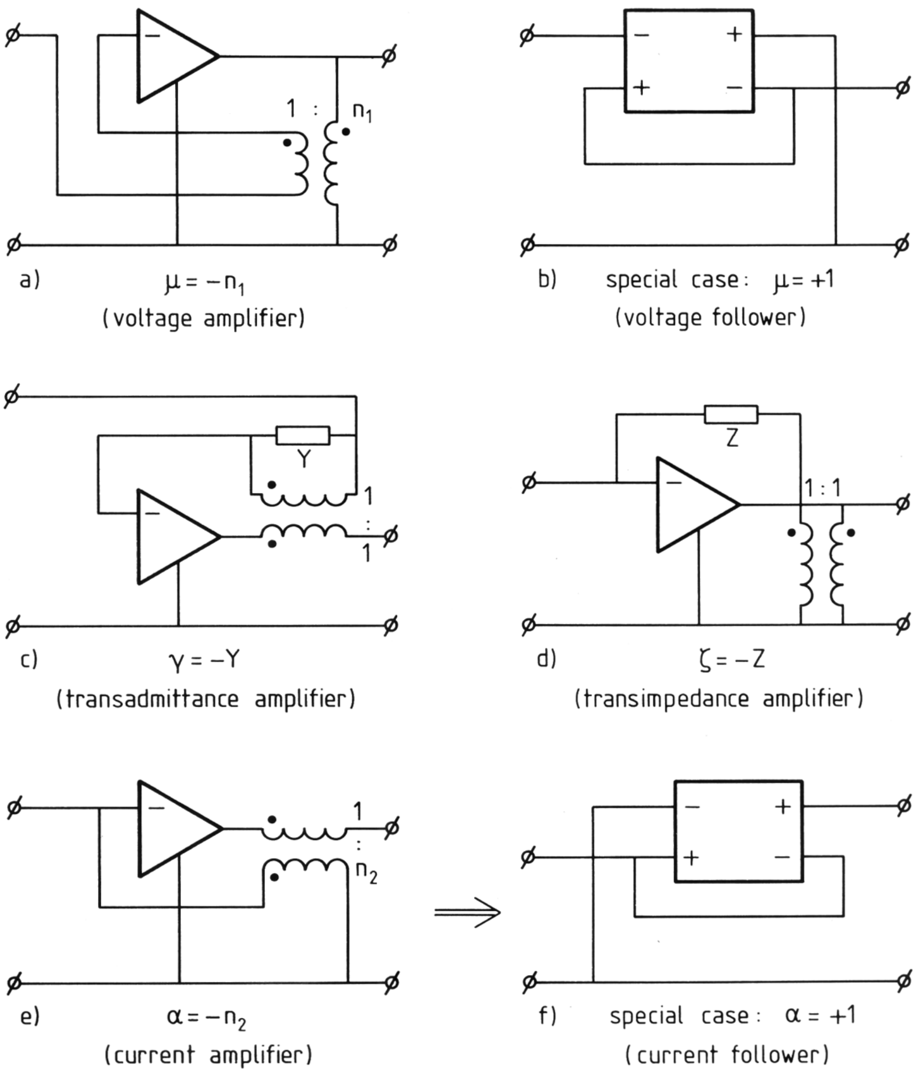


Figure 1.14 *Single-loop amplifiers with zero or infinite input and output impedances.*

(iii) Dual-loop configurations

The amplifiers with two transformer-feedback loops form a group that — to the author's knowledge — has not been classified as such before. Isolated examples can be found in literature [12], [13], [14].

Non-energicness is violated in all cases where impedances are used. The deterioration of SNR and PE is determined by the types and the magnitudes of the impedances in the feedback networks, together with the turns ratios of the transformers. Large values of Z and Y are advantageous, as will be shown in chapters 3 and 4. The configurations are able to accomplish characteristic impedance matching at input and/or output, just like the non-energic types presented in section 1.4.2.

We will discuss briefly the properties of the configurations that follow directly from figure 1.12 by making two out of the four quantities Y, Z, n_1 and n_2 infinite. Some alternative configurations, based on the voltage and current followers of figure 1.14b and f, are presented in addition.

– The four configurations with either an accurate input or an accurate output impedance are given in figure 1.15. Their transfer properties correspond to configuration numbers 6, 7, 0 and 11 from table 1.3. They are essentially unilateral. For obtaining these specific transfer and impedance properties with passive feedback networks, transformers are indispensable. Only the transimpedance parameters can be fixed without a transformer, provided that the active part is inverting. Non-energicness is lost as a result of the use of impedances in the feedback networks. By cascading the appropriate configurations, characteristic-impedance matching at both input and output can be realised.
– The two non-unilateral configurations with transfer properties corresponding to the configuration numbers 8 and 9 in table 1.3 are given in figure 1.16. Figure 1.16a shows the two-transformer configuration which has essentially non-energic feedback. Figure 1.16b represents an approximation of the two gyrator configuration. Alternative dual-loop non-unilateral configurations with transfer properties similar to the two-transformer configuration of figure 1.16a can be obtained with the follower circuits of figures 1.14b and f in combination with only one transformer. The gain factors of these followers can be raised above unity with the aid of auto-transformers so that a voltage amplifier and a current amplifier result, as shown in figure 1.17a and b, respectively.

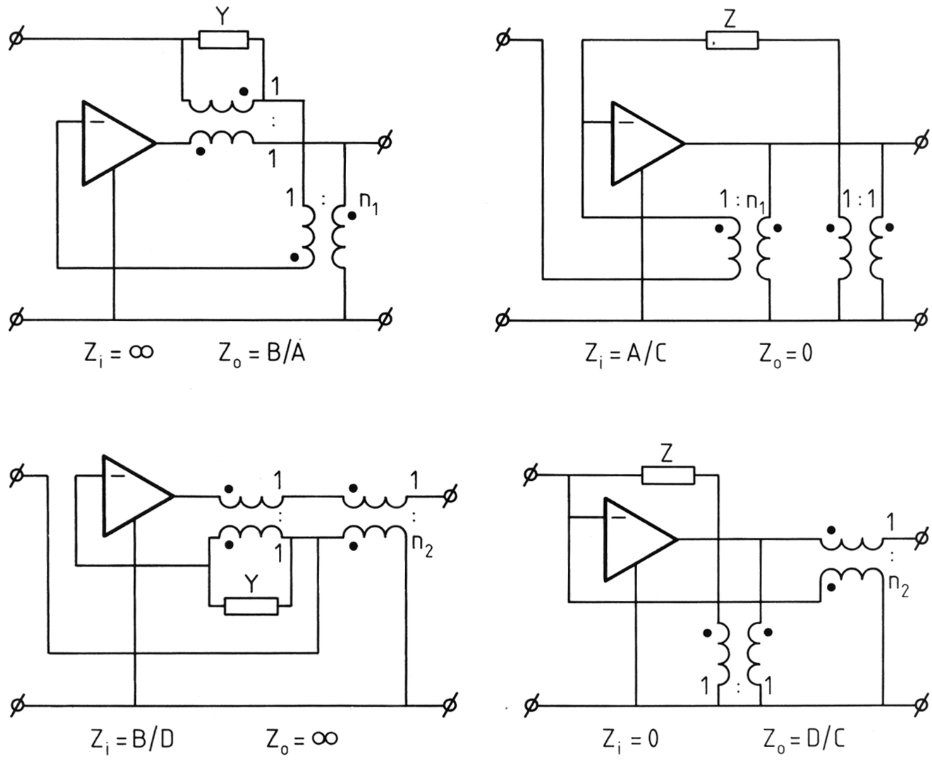


Figure 1.15 Dual-loop configurations with accurate impedance at one port and either zero or infinite impedance at the opposite port.

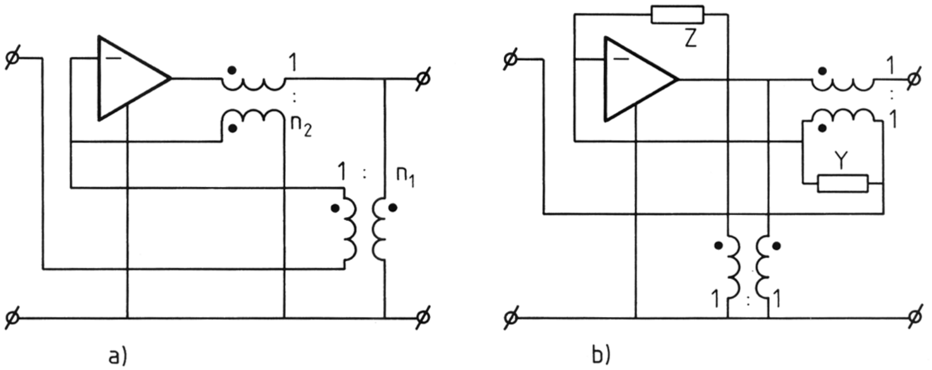


Figure 1.16 Non-unilateral dual-loop configurations.

By providing the transformers with secondary windings, the output current in the circuit of figure 1.17a can be sensed and a part of it can be fed back. In figure 1.17b the same can ho done with the output voltage. Figure 18 shows these extensions. The circuit of figure 1.18d is described in a patent by Norton [14], where the active part is formed by a single CB stage. All configurations of figure 1.18 have essentially non-energic feedback, but suffer from the disadvantage of parasitic effects duo to the floating ports of the active part. Moreover, matching errors at one port are reflected at the other port.

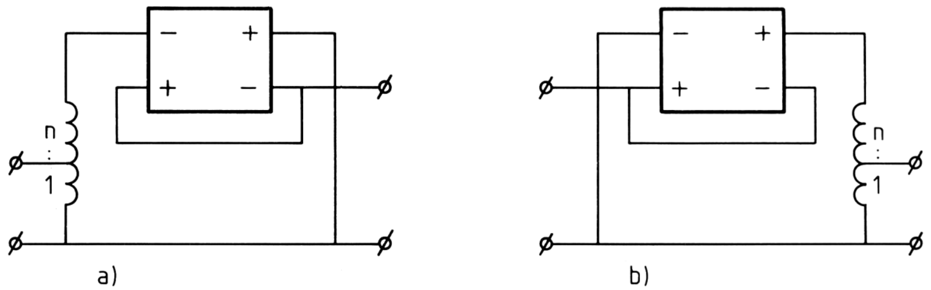


Figure 1.17 Voltage-gain factor μ and current-gain factor α of the follower configurations enlarged to a value n with the aid of auto-transformers.

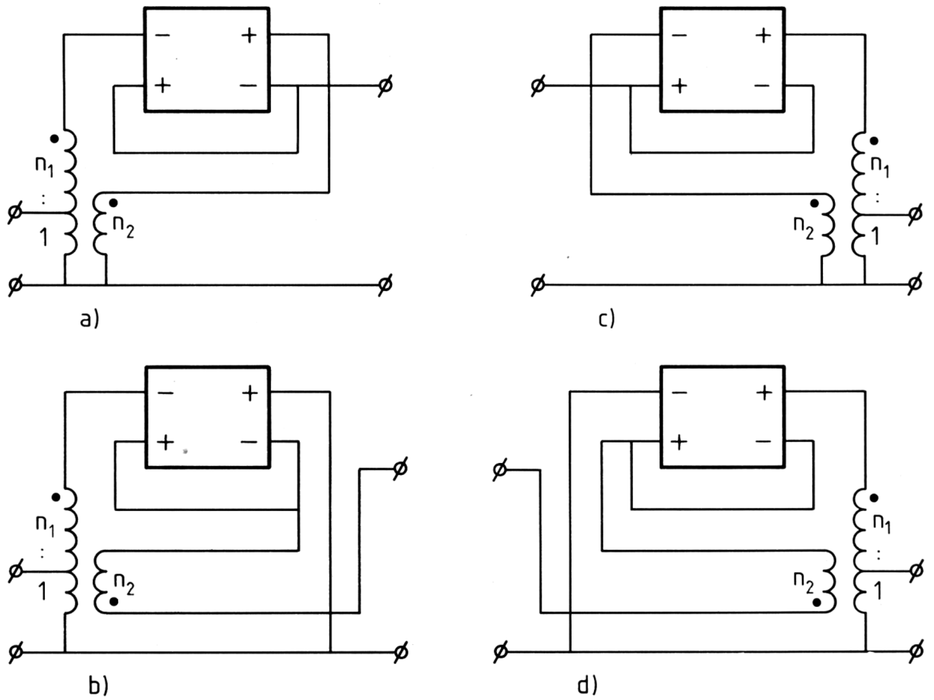


Figure 1.18 Alternative configurations for non-energic transformer feedback, using followers and auto-transformers.

(iv) Three-loop configurations

Realisation of three-loop amplifiers will not be discussed because of their small practical significance.

(v) Four-loop configuration

The parameters of the four-loop configuration are given by fairly complex expressions. To get an impression of the properties of this amplifier, some reasonable approximations will therefore be made, namely: $n_1 \gg 1$, $n_2 \gg 1$, $YZ \gg 1$.
The parameters B and D are determined with a short-circuited output, i.e. $U_o = 0$. With the assumption that $I_i \ll I_o$, we find:

$$B \approx \frac{-1}{Y}, \quad D \approx -(1/n_2 + 1/YZ)$$

The parameters A and C are determined with an open output; i.e. $I_o = 0$.
We find:

$$A \approx -(1/n_1 + 1/YZ) \quad \text{and} \quad C \approx -1/Z$$

A unilateral transfer is possible when the condition $AD = BC$ is met, or when

$$Y^2Z^2 + (n_1 + n_2 - n_1n_2)YZ + n_1n_2 = 0$$

which reduces with the given condition to $YZ = n_1n_2$. We will not work out the possibilities of this rather complicated feedback configuration any further.
We will conclude this section by referring to a circuit having four feedback loops around one active device which was patented in 1971 [15]. By using an active part with floating ports (one transistor), two transformers can be omitted. The circuit mentioned has equal input and output impedances of 50 Ω, quite insensitive to errors in the source and load impedances. The transformer feedback is realised with a directional coupler. Figure 1.19 shows the basic circuit diagram.

1.4.4 Realisation of transmittances without the use of transformers

There are many practical situations where the use of transformers cannot be tolerated, either because of their loss favourable electrical properties, especially at low frequencies, or for economical reasons. We therefore have to look for other realisations of the feedback networks using impedances exclusively. For reasons of interference rejection it is generally required that source, load and amplifier have one common terminal. In that case (the only one to be dealt with here), equipping the active part with floating ports according to figure 1.20 for current sensing at the output and for voltage comparison at the input cannot be avoided.

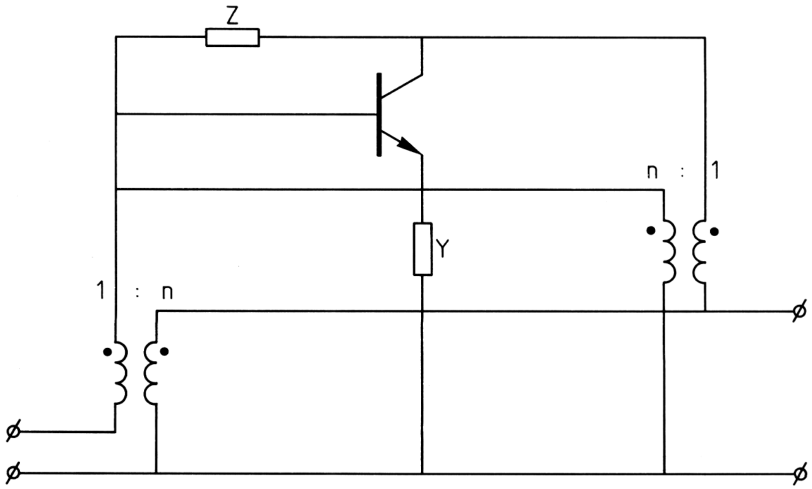


Figure 1.19 Amplifier with four feedback loops using a directional coupler.

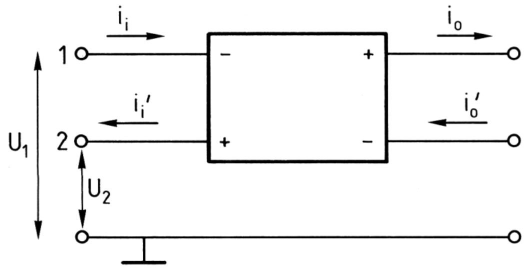


Figure 1.20 Active part with ports floating with respect to ground.

For perfect current sensing, the currents i_o and i'_o (figure 1.20) must be exactly equal. For a perfect comparison of the source voltage and the feedback voltage, the common-mode rejection of the active part must be infinitely large. In other words: the voltage sensitivities of both input terminals must be exactly equal, but must have opposite signs.

The ideal situation cannot, of course, be realised in practice. The active part is connected via finite impedances to a power supply and has finite physical dimensions. Therefore, it is either directly or via parasitic impedances connected to ground, and the common-mode rejection cannot be infinitely large. These imperfections in the active part lead to departures from the ideal asymptotic-gain value. We will ignore them at this stage and account for them in chapter 4.

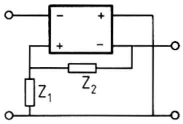
The configurations that can be realised with the aid of active parts with floating ports and passive components, except transformers, are given in figure 1.21. The number of basically different realisations is reduced to only six. Some special cases of the voltage and current amplifier and of the two-loop configurations are added. All other configurations listed in table 1.2, section 1.4.2 require inverting feedback networks, which can either be realised with transformers or with active devices. We dealt with transformers in the preceding section. Active feedback networks will be treated concisely in the next section.

The values of the parameters of the basic configurations are given in the figure captions in figure 1.21. Only the configurations with unity-gain factors are non-energic. The other configurations have impedances in series or in parallel with the signal path, so that signal-to-noise ratio and efficiency are somewhat deteriorated as will be shown in chapters 3 and 4.

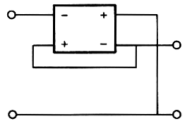
Although the series impedances can have small values, their effect on input or output impedances can be large. Shunt impedances may be largo and still have a great effect on impedances as well. In the transadmittance and transimpedance amplifiers, the influences of the series and shunt impedances on the signal-handling capability and signal-to-noise ratio are fixed if the transfer functions must have certain prescribed values. Compromises are possible in the voltage and current amplifiers, where for optimum noise performance Z_1 must be small and Z_2 must be large, respectively. For maximum efficiency, Z_2 must be large and Z_1 must be small. For given values of the voltage or current gain these requirements may be conflicting, so that efficiency must be sacrificed in favour of signal-to-noise ratio, or vice versa.

Lossless impedances can sometimes be used in the feedback paths. Though they produce no noise, they can have an adverse influence on the signal-to-noise ratio and though they dissipate no signal power, signal handling may decrease. In chapters 3 and 4 we will pay more attention to these aspects.

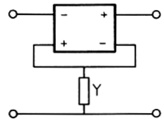
The first six configurations in figure 1.21 have single feedback loops. When loop gain is infinite, the input and output impedances are either zero or infinite. Their transmittances are completely determined by the passive components of the feedback networks. The remaining configurations have two feedback loops and are capable of realising accurate input and output impedances, provided that the source and load impedances are accurately known. The features of the two-loop configurations will be studied in some detail.



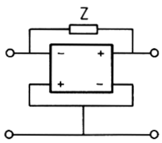
a. $\mu = +(1+Z_2/Z_1)$
(voltage amplifier)



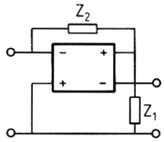
b. special case: $\mu = 1$
(voltage follower)



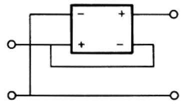
c. $\gamma = -Y$
(transadmittance amplifier)



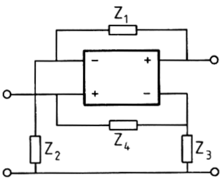
d. $\zeta = -Z$
(transimpedance amplifier)



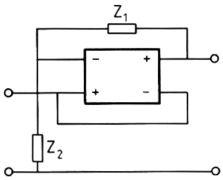
e. $\alpha = +(1+Z_2/Z_1)$
(current amplifier)



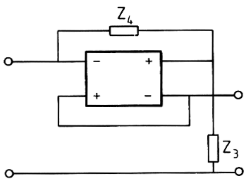
f. special case: $\alpha = +1$
(current follower)



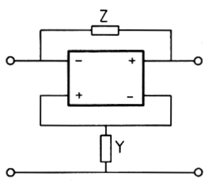
g. $\mu = (Z_2+Z_1)/Z_2$, $\gamma = \infty$
$\zeta = (Z_1+Z_2)(Z_3+Z_4)/(Z_2+Z_3)$,
$\alpha = (Z_3+Z_4)/Z_3$



h. special case: $\mu = (Z_1+Z_2)/Z_2$, $\gamma = \infty$
$\zeta = Z_1 + Z_2$, $\alpha = 1$



i. special case: $\mu = 1$, $\gamma = \infty$
$\zeta = Z_3 + Z_4$, $\alpha = (Z_3+Z_4)/Z_3$



j. $\mu = 1 - YZ$, $\gamma = (1-YZ)/Z$
$\zeta = (1-YZ)/Y$, $\alpha = 1 - YZ$

Figure 1.21 *Amplifier configurations with passive feedback components with the exception of transformers.*

(i) Two-loop configuration of figure 1.21g

Configuration 1.21g has the following transmission parameters:

$$A = \frac{Z_2}{Z_1 + Z_2}, \quad B = 0, \quad C = \frac{Z_2 + Z_3}{(Z_1 + Z_2)(Z_3 + Z_4)}, \quad D = \frac{Z_3}{Z_3 + Z_4}$$

It has the characteristics, therefore, of a three-loop amplifier as discussed in section 1.4.2.

With

$$Z_i = \frac{AZ_\ell}{CZ_\ell + D} \quad \text{and} \quad Z_o = \frac{DZ_s}{A + CZ_s}$$

it follows that an accurate input impedance can be realised, provided that the load impedance Z_ℓ is accurately known, or infinitely large. An accurate output impedance is obtained when the source impedance Z_s is accurately known or infinitely large. Characteristic matching at both the input and the output is not possible in an accurate way, unless a series impedance is inserted in the input or output leads in order to fix parameter B.

The configuration is suitable, for example, for realising an accurate and linear low-noise damping resistance for magneto-dynamic transducers [16]. Moreover, a frequency-dependent transfer function can be realised at the same time [17]. The circuit is frequently used for impedance matching at input and output [18] in spite of its inferiority compared with the circuit of figure 1.21j.

(ii) Two-loop configuration of figure 1.21j

The transmission parameters of the configuration in figure 1.21j are given by:

$$\mathrm{A} = \frac{1}{1 - YZ}, \qquad \mathrm{E} = \frac{Y}{1 - YZ}$$

$$\mathrm{B} = \frac{Z}{1 - YZ}, \qquad \mathrm{D} = \frac{1}{1 - YZ}$$

Though all parameters have accurate values, the configuration does not have the versatile properties of the ideal four-loop amplifier. The condition for realising input and output impedances that are independent of source and load impedances coincides with a zero transfer condition. In other situations, the configuration is not unilateral.

Characteristic impedance matching at both sides is possible under the condition:

$$R_c^2 = \frac{Z}{Y},$$

where $R_c = R_i = R_s = R_o = R_\ell$.

The transfer function can be frequency dependent at the same time if Z and Y have the same frequency dependency. The circuit is, in principle, more suited to realising input and output impedance matching than the configuration of figure 1.21g. Nevertheless, it is probably used less for this purpose [19], [20].

1.5 Realisation of transmittances with active feedback networks

1.5.1 Introduction

In the previous sections we have seen that a number of amplifier types can only be realised with the aid of inverting feedback elements. The only passive feedback element that can be used in practice for this purpose is the transformer. In order to enable the designer to realise, for example, inverting current amplifiers, or amplifiers with an accurate impedance at one port and a zero or infinite impedance at the other, without the use of transformers, we have to resort to active feedback networks that have to accomplish the required inversion.

It will appear to be possible and meaningful to distinguish two fundamentally different methods for realising negative feedback with active feedback networks. One will be referred to in the following as the *active-feedback* technique. It is similar to the passive-feedback technique in so far as the output quantity itself is sensed and the feedback quantity is compared directly with the input quantity. The other, which will be called an *indirect-feedback* technique, does not — when applied to the output — sense the output quantity itself, but instead a quantity which resembles the output quantity as much as possible. An indirect comparison of the input quantity with the feedback quantity is equally possible. The indirect-feedback method thus bears some resemblance to pantographic techniques.

It should be noted that the active- and indirect-feedback techniques are basically inferior to passive-feedback techniques with respect to nearly all quality aspects. The configurations are interesting in the first place because they supplement the number of transmittance realisations. Moreover, some of them are especially attractive for implementation as integrated circuits.

1.5.2 Indirect-feedback amplifiers

The use of an indirect-feedback technique may be considered when a transmittance with a sign opposite to that obtained in the corresponding passive-feedback configuration is needed, or when an accurate input or output impedance at one port of the amplifier is required.

Indirect feedback at the output of an amplifier differs from the direct-feedback methods discussed up to now in so far as the actual quantity supplied to the load is not sensed. Instead of sensing the actual load current or voltage, the output quantity of a dummy output stage is sensed, and the load quantity is supplied by a similar second stage. The transfer function of this actual output stage is not enclosed by the feedback loop. For obtaining an accurate and linear amplifier transfer, certain parameters of both output stages have to be equal. Furthermore, their relevant transfer characteristics have to be linear or equally non-linear.

Indirect-feedback techniques at the input can be described in a similar way. The examples that will be given below are self-evident.

The indirect-feedback techniques to be discussed may not be basically now. They are similar or at least related to techniques described earlier [21], [22]. The explicit formulation and generalisation of these techniques as feedback techniques, however, is believed to be justifiable because it may be useful in various amplifier design problems.

We will not discuss all possible types of indirect-feedback configurations, but will instead confine ourselves to examples of configurations that yield the most interesting transfer properties from the viewpoint of utility.

(i) The technique of indirect current sensing

Figure 1.22 shows, as an example, the principle of indirect current sensing combined with direct current comparison (shunt feedback) at the input. Two equivalent output stages I and II are needed. The output current I_o of stage I instead of the actual output current I_ℓ is sensed, and part of this current is fed back to the amplifier input. The bias conditions in the output stages are assumed to be equal. The transfer properties of the output stages are characterised by their transmission parameters, and the loop gain in the direct-feedback loop (formed by the active part AP, output stage I and the current divider Z_1-Z_2) is assumed to be infinite. In that case U and I are necessarily zero and the asymptotic value of the current gain is given by:

$$A_{i\infty} = \frac{I_\ell}{I_s} = -\left(1 + \frac{Z_2}{Z_1}\right)\left(\frac{AZ' + B}{A'Z_\ell + B'}\right) \tag{1.28}$$

where $$Z' = \frac{Z_1 Z_2}{Z_1 + Z_2}$$

The parameters C and D of the output stages obviously don't play any role in the transfer function in this ideal situation. Effects of a finite transimpedance ($1/C$) and of the current-gain factor ($1/D$) can therefore be reduced by increasing the loop gain in practical situations.

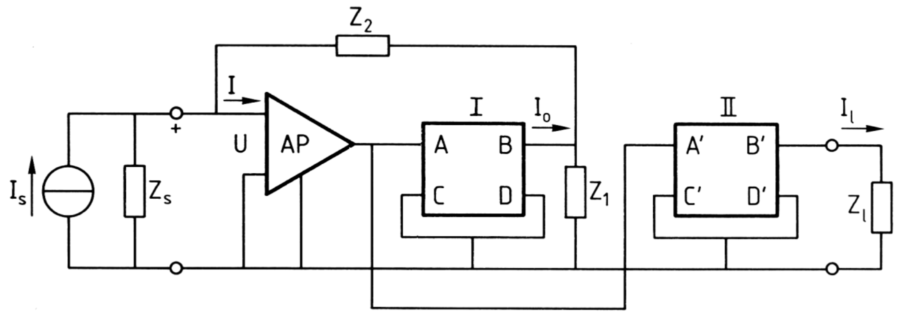


Figure 1.22 Current amplifier with indirect current sensing.

In a direct-feedback amplifier, the stage parameters need not influence the transfer function when the loop gain is infinite. In this indirect-feedback situation, however, output-stage parameters A and B occur in the transfer function of the amplifier, and because their values are signal dependent, linearity is not guaranteed even when the loop gain is infinite. Special measures for the reduction of this non-linearity may be necessary. The ratio of the parameters A and B represents the output impedance of a stage under voltage-drive conditions. The non-linearity can therefore alternatively be ascribed to the signal-induced variations of this output impedance.

For obtaining perfect linearity, it is necessary that signal voltages and currents in both stages be equal. By meeting the condition that $Z' = Z_\ell$ follows from (1.28) that the transfer function is determined by the passive impedances Z_1 and Z_2 exclusively because, in that case, A and A' as well as B and B' can be mutually equal, resulting in a perfect compensation. An obvious disadvantage of meeting this condition, however, is that both output stages have to handle equal signal powers and, in fact, an output-power loss of 3 dB results. Moreover, the value of Z_ℓ must be well known.

The power loss can be reduced by using several identical configurations in parallel in stage II. When n configurations are paralleled in stage II, the parameter A' remains unaltered, whereas the parameter B' is reduced by a factor n, provided that the bias conditions in all configurations are equal. When the output stages can be implemented so that the influence of the terms with A and A' can be disregarded, the current gain of this amplifier becomes:

$$A_{i\infty} = \frac{I_\ell}{I_s} = -n(1 + Z_2/Z_1)$$

This current transfer is linear provided that the transconductances of all output configurations vary in the same way along with the signal. The signal power handled by output stage I can thus be significantly smaller than that handled by output stage II. Large amounts of output power loss are avoided.
Scaling the gain in accordance with the number of paralleled output stages offers the possibility of realising indirect-feedback configurations with arbitrary current gains without using any passive impedances. A very simple example is given in figure 1.23, where CE-configurations are used as output stages. The configuration with an inverting *unity current gain* is generally known as a *current mirror*.

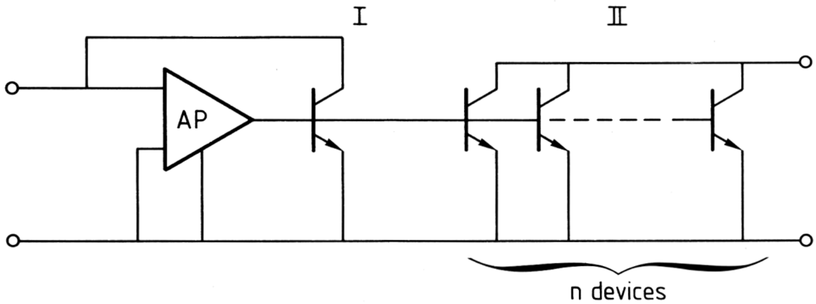


Figure 1.23 Current amplifier with gain scaling achieved by using n paralleled devices in stage II.

(ii) Output-stage implementation
The question arises, as a matter of course, of how the output stages of figure 1.22 can be implemented so that the influence of transmission parameter A can be disregarded ($AZ_\ell \ll B$). It will be clear that we have to be content with approximations that will be valid only in cases where the load impedances are relatively small. A single transistor in a common-emitter or common-source configuration, as in figure 1.23, may be a reasonable choice when the load impedance is low, but more elaborate output stages, whether or not with negative feedback, may give much better results.
At this stage it is appropriate to pay some attention to the unique capabilities of common-emitter differential output stages. Figure 1.24 shows an indirect-feedback current amplifier equipped with such stages. When biased at equal tail currents, the circuit behaves like a unity-gain current amplifier with both an inverting and a non-inverting output. Thanks to the exponential relationship between base-to-emitter voltage and collector current in *a bipolar* transistor (or in an MOS transistor in the

subthreshold region), and consequently, to the fact that its transconductance is linearly proportional to the collector current, the current gain of this amplifier can be varied by varying the ratio of the tail currents. For small load impedances ($AZ_\ell \ll B$), the current transfer characteristic remains linear, independent of this ratio. The circuit of figure 1.24 can alternatively he considered to operate according to the *translinear principle,* as first formulated by Gilbert [23]. Such a point of view, however, fits less well in a classification of possible negative-feedback configurations.

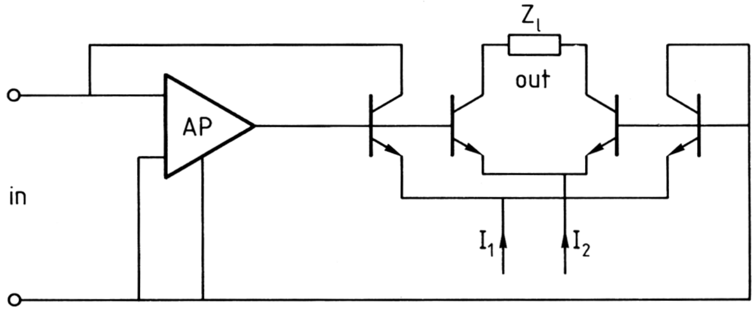


Figure 1.24 Indirect–feedback current amplifier with differential output configurations.

Further aspects of output-stage implementation will be put off until later chapters, where the influence of the output stage parameters on transfer linearity and accuracy will be studied in more detail.

(iii) Indirect current sensing combined with input series feedback

Instead of realising low input impedances with the aid of input shunt feedback, as in the configurations discussed up to this point, it is equally possible to realise high input impedances by means of input series feedback. A basic configuration of such a transadmittance amplifier is shown in figure 1.25. The active part must have a floating input port, just as in conventional transadmittance amplifiers. Assuming an ideal active part, the asymptotic value of the transadmittance is given by:

$$Y_\infty = \frac{I_\ell}{U_s} = Y\left(\frac{A/Y + B}{A'Z_\ell + B'}\right) \qquad (1.29)$$

The factor in parentheses has the same form as in expression (1.28), and the discussions in connection with that expression apply equally well to this configuration.

A noticeable feature of an indirect-feedback transadmittance amplifier is that it combines a high input impedance with the possibility of obtaining a variable gain by using differential output configurations similar to those of the indirect-feedback current amplifier of figure 1.24. This opens up the possibility of using an additional passive feedback loop in order to realise an accurate and linear input impedance that can he controlled by varying the ratio of the currents in the output stages. Such a current-controlled impedance may he useful, for example, for automatic gain control in oscillators or in filters.

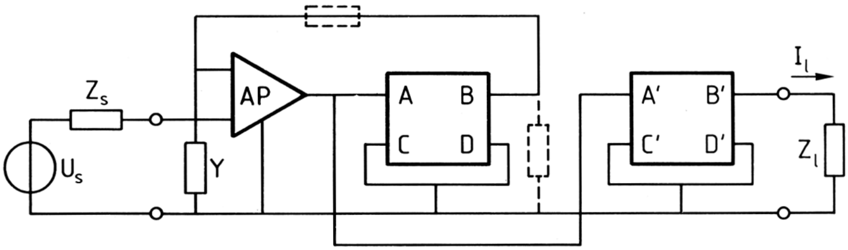


Figure 1.25 Indirect-feedback transadmittance amplifier.

(iv) The technique of indirect voltage sensing

A technique similar to that described above might he used to realise amplifiers with indirect voltage sensing for fixing either the transmission parameter A or C. however, the sign of the voltage-gain factor or transimpedance is the same as that of an amplifier with direct voltage sensing, unless equivalent inverting and non-inverting output stages are available. These output stages should have zero output impedances in the ideal case. This desired combination of properties cannot be easily obtained. Fortunately, there seems to be no urgent need for these amplifier types.

(v) The technique of indirect voltage comparison

To illustrate the operating principle of amplifiers with an indirect comparison of the feedback voltage and the signal-source voltage, a configuration with *output shunt feedback* is shown in figure 1.26.

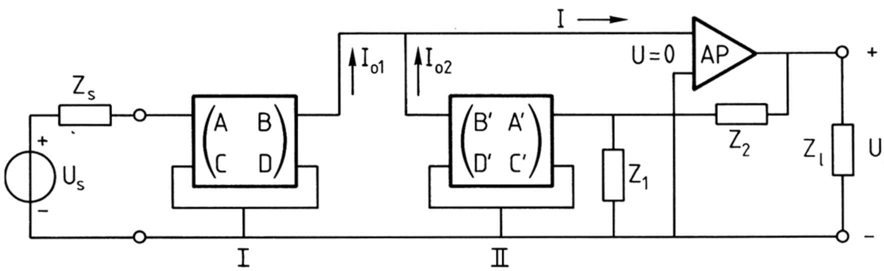


Figure 1.26 Inverting voltage amplifier with indirect voltage comparison.

Two input stages are needed. Their transfer properties are described again with the transmission parameters. Assuming the loop gain in the direct-feedback loop (formed by input stage II, the active part AP and the voltage divider Z_1-Z_2 to be infinite, the voltage U and the current I are necessarily zero. The asymptotic value of the voltage gain then becomes:

$$A_{u\infty} = \frac{U_\ell}{U_s} = -(1 + Z_2/Z_1)\left(\frac{B' + D'Z'}{B + DZ_s}\right) \tag{1.30}$$

where $$Z' = \frac{Z_1 Z_2}{Z_1 + Z_2}$$

Note that the parameters A and C (characterising the voltage-gain factor and the transimpedance, respectively) ideally do not play any role in the amplifier transfer. Unlike direct-feedback amplifiers, but like the technique with indirect current sensing, two parameters (B and D) of the input stages influence the transfer function. Because these parameters are signal dependent, the transfer function is non-linear even when loop gain is infinite. For obtaining satisfactory linearity compensation techniques may be necessary.

The ratio of the parameters B and D represents the input impedance of the input stages with short-circuited output. Non-linearities can therefore alternatively be ascribed to the signal-induced input impedance variations.

A necessary condition for obtaining perfect linearity by compensation is that $Z' = Z_s$ The currents i_{o1} and i_{o2} have equal magnitudes, but opposite signs. As a consequence, either *linear, symmetrical or complementary* transfer characteristics of the input stages are additionally required. The parameters B and B' as well as D and D' then can have equal voltage dependencies. when both conditions are met, the voltages u_o and u_s have the same forms, but opposite signs, and a linear and accurate voltage amplifier results.

An obvious disadvantage of meeting the condition $Z' = Z_s$ is that both input stages equally contribute to the total amplifier noise. The amplifier-noise contribution is therefore 3 dB higher than it would be when determined by stage I exclusively. A second disadvantage is that the source impedance must be well known. The noise contribution of stage II can be reduced, by reducing its transadmittance by connecting several devices in series.

A general configuration with the equivalent input noise sources of stages I and II is given in figure 1.27. The parameter D of stage II does not change significantly by the series connection, but the parameter B is enlarged by a factor n, provided that the bias conditions in all stages are equal.

The voltage gain of the configuration of figure 1.27 is given by:

$$A_{u\infty} = \frac{U_\ell}{U_s} = -n(1 + Z_2/Z_1)$$

provided that the influence of the parameter D can be disregarded. When the power-density spectrum of the input noise voltage u_{nI} of stage I is denoted as $S(u_{n\mathrm{I}})$, stage II will have an input noise voltage characterised by a spectrum about n times as large: $S(u_{n\mathrm{II}}) \approx nS(u_{n\mathrm{I}})$. The input noise current sources $i_{n\mathrm{I}}$ and $i_{n\mathrm{II}}$ have equal spectra. The spectra of the output noise currents of the stages due to the input noise voltages are given by:

$$S(i_{o\mathrm{I}}) = \frac{S(u_{n\mathrm{I}})}{B^2} \quad S(i_{o\mathrm{II}}) = \frac{nS(u_{n\mathrm{I}})}{n^2B^2} = \frac{S(u_{n\mathrm{I}})}{nB^2}$$

The relative contribution to the total equivalent input noise voltage is therefore smaller as n increases. The influence of the noise current source $i_{n\mathrm{II}}$ can be kept small by taking a small value for Z'.

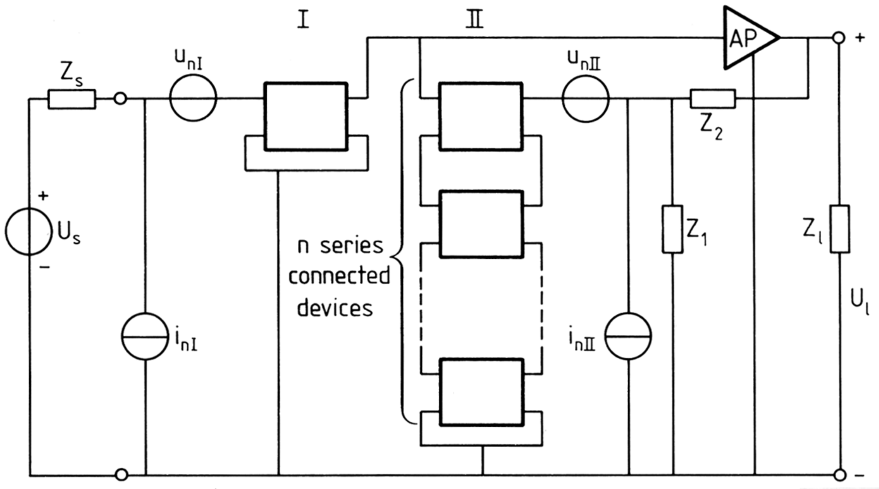


Figure 1.27 Indirect-feedback voltage amplifier with equivalent input noise sources.

We can conclude that, when the gain scaling is achieved by series connection of configurations, noise performance need not significantly be deteriorated. The limits of the amplifier noise performance can thus be almost completely set by the noise properties of stage II (provided that the noise contribution of the active part is small).

(vi) Input-stage implementation

Two simple input-stage implementations are shown in figure 1.28. Complementary configurations are used in figure 1.28a, whereas the differential pairs of figure 1.28b have symmetrical transfer functions. The latter configuration can be made to have an inverting as well as a non-inverting input.

With small values of the source impedance and of Z', the terms with D and D' in (1.14) may be sufficiently small to be disregarded. The remaining non linearity may arise from unequal variations in the stage transconductances $1/B$. It should be noted that unequal bias currents in the devices generally result in non-linearity. For maintaining linearity it would be necessary for the transadmittance of the stages to be linearly proportional to the driving input voltages. In bipolar transistors this is not the case, but when field effect devices are used in the input stages, linearity is hardly harmed by unequal currents because the above condition is roughly met. FET input stages are favourable also because the parameter D is much smaller, at least at low frequencies. Linearity can be improved alternatively by applying, for example, negative feedback to the input stages. Further considerations with respect to input-stage implementation will be put off until later chapters.

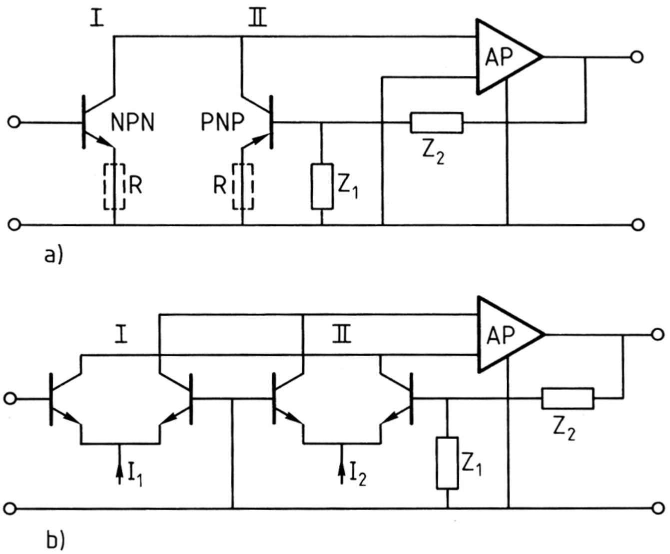


Figure 1.28 Simple implementations of input stages in indirect-feedback voltage amplifiers. Configuration a) has complementary input stages. Configuration b) has symmetrical input stages.

(vii) Indirect voltage comparison combined with output series feedback

Instead of realising low output impedances by using output shunt feedback, it is equally possible to obtain high output impedances by means of output series feedback. A basic configuration of such a transadmittance amplifier is shown in figure 1.29. The active part AP must have a floating output port, the same as in conventional transadmittance amplifiers. Assuming an ideal active part, the transadmittance of the amplifier becomes:

$$Y_\infty = \frac{I_\ell}{U_s} = Y\left(\frac{B + D/Y}{B + DZ_s}\right) \tag{1.31}$$

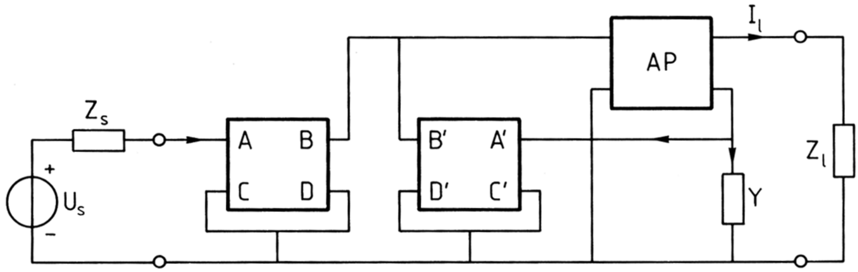


Figure 1.29 Transadmittance amplifier with indirect voltage comparison.

The parenthesised factor has the same form as in expression (1.30), and the discussions with respect to that expression apply equally well to this configuration.

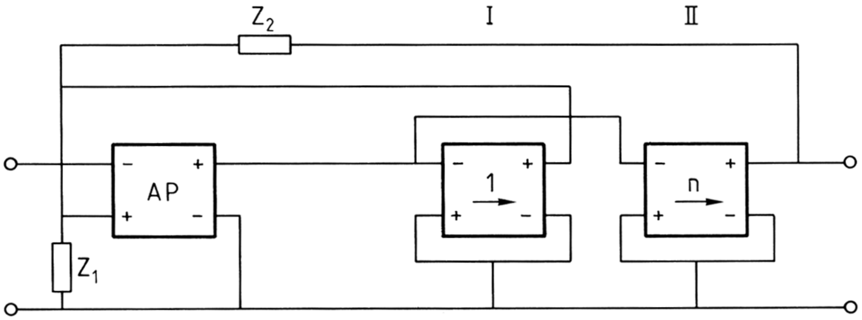
(viii) The technique of indirect current comparison

A technique similar to that described above might be used to realise amplifiers with indirect current comparison, fixing either the parameter C or D. However, the signs of the transimpedance or current-gain factor thus obtained are equal to those of the corresponding direct-feedback configurations, unless equivalent inverting and non-inverting input stages with very low input impedances are available. These properties are difficult to obtain but there seems to be no immediate need for these configurations.

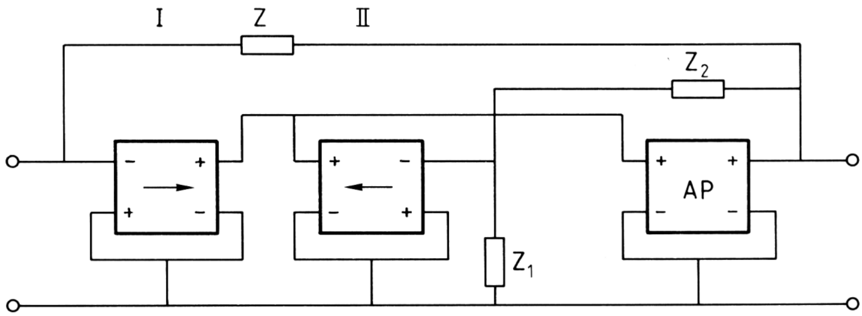
(ix) Dual-loop indirect-feedback configurations

In the foregoing we have shown that it is possible to realise inverting accurate and linear current- and voltage-gain factors by means of indirect feedback as well as non-inverting transadmittances. By fixing a second transfer parameter of such indirect-feedback configurations by means of a direct-feedback loop, it becomes possible to obtain an accurate input or output impedance at one port in combination with a zero or infinite impedance at the other port. Table 1.3 in section 1.4.2 shows the various combinations of transmission parameters that have to be fixed for this purpose. Figure 1.30 shows basic configurations of amplifier types 6, 7, 10 and 11.

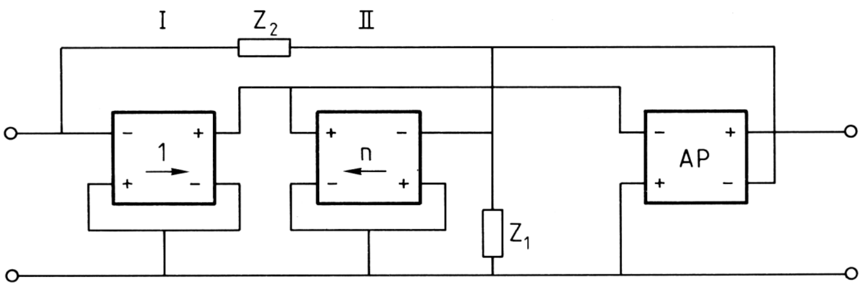
The arrows indicate the directions of the signal transfer. The examples are believed to be self evident.



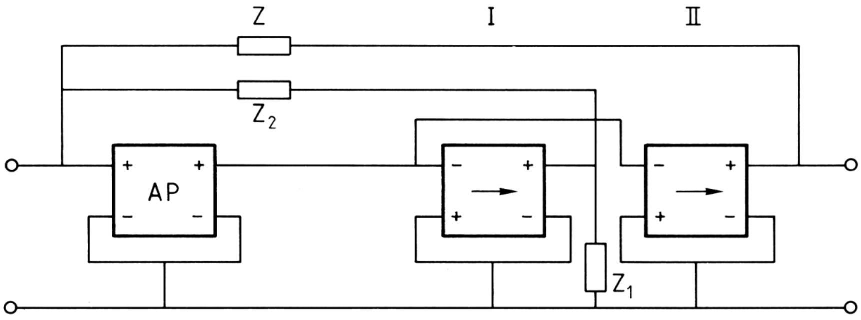
(a) realisation of configuration 6 $B = Z_1/n$ $A = 1/(1 + Z_2/Z_1)$



(b) realisation of configuration 7 $A = -1/(1 + z_2/Z_1)$ $C = -1/Z$



(c) realisation of configuration 10 $B = Z_1/n$ $D = 1/(1 + Z_2/Z_1)$



(d) realisation of configuration 11 $C = -1/2$ $D = -1/(1 + Z_2/Z_1)$

Figure 1.30 Indirect-feedback configuration with an additional direct-feedback loop for the realisation of dual-loop amplifier types.

In the foregoing it has been demonstrated that indirect-feedback techniques offer useful possibilities for realising amplifier configurations that cannot be obtained with conventional (no transformers) passive feedback networks. The active-feedback techniques to be discussed in the next section offer similar possibilities. They may yield better approximations, however, of the ideal situation, where the transmittance is determined by passive elements exclusively.

1.5.3 Active-feedback amplifiers

The use of an active-feedback technique may be considered when a transmittance with a sign opposite that of the corresponding passive-feedback configuration is needed. It can serve as an alternative for an indirect-feedback configuration. This technique also offers the possibility to realise the dual-loop amplifiers 6, 7, 10 and 11 of table 1.3, section 1.4.2. We will not go into much detail in dealing with the possible configurations, but will only give some examples.

Figure 1.31 illustrates the principle of operation of an active-feedback configuration. The transmission parameter D has been fixed in this case by using an inverting current attenuator as a feedback element. In order to realise an amplifier with a linear and accurate transfer function, the inverting current attenuator itself should be a negative-feedback circuit, either with direct feedback or with indirect feedback.

In section 1.4.4 we have seen that it is not possible to realise an approximately non-energic negative-feedback inverting current amplifier (or attenuator) while using a passive feedback network. Making an inventory of possible configurations of the feedback network, we find:

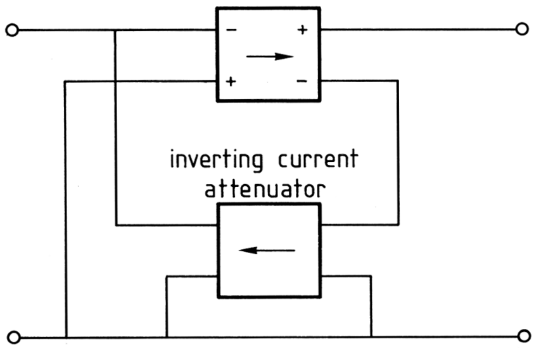


Figure 1.31 Inverting current amplifier with active feedback.

- A transimpedance amplifier with a series impedance at its output;
- A transadmittance amplifier with a shunt impedance at its input;
- An indirect-feedback current amplifier (with a current-gain factor smaller than unity).

The first two configurations are shown in figure 1.32. Assuming nullor properties of the active parts in both the forward path and in the feedback network, the current-gain factors of both amplifiers have values $\alpha = -Z_2/Z_1$.

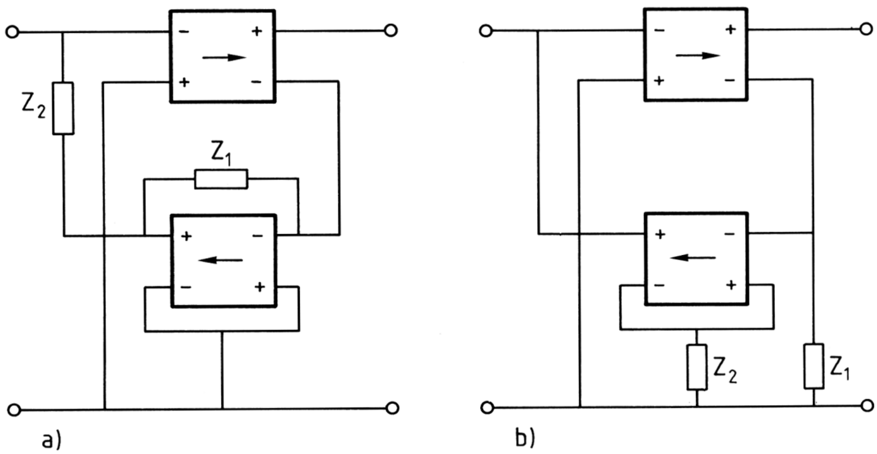


Figure 1.32 Two inverting current amplifiers with active feedback. $\alpha = -Z_2/Z_1$.

Though the active feedback networks basically have a non-linear transfer function, their influence on amplifier non-linearity may be rather small when the current and voltage variations at the output of the feedback network remain small. The output voltage is kept small by taking a small Z_1. For optimum noise performance, however, the impedances Z_2 should have values as large as possible (as will be elucidated in chapter 3). A compromise between noise and distortion performance

may therefore be necessary.
Raising the loop gain in the active part of the active feedback network may, of course, improve its linearity. Stability considerations will finally be decisive for its allowable complexity.
Similar considerations can be given for other single-loop active-feedback configurations. An important criterion for their usefulness is the extent to which they behave like approximations of non-energic feedback amplifiers. In other words, the question is can they perform better than ‘brute-force’-terminated amplifiers, where series and/or shunt impedances at the input or output are used? Suffice it to note here that they can, under certain conditions. We will address this question in some more detail in chapters 3 and 4.
The active-feedback technique enables us to realise inverting voltage and current amplifiers and non-inverting transimpedance and transadmittance amplifiers. In addition, there is a possibility to realise amplifiers with an accurate impedance at one port and an (ideally) infinite or zero impedance at the other port. Figure 1.33 shows some basic configurations, which can be found with the aid of table 1.3, section 1.4.2. The arrows within the active-part symbols indicate the directions of the signal transfer. The active feedback networks in figure 1.33 are realised with the basic configurations of figure 1.21. Alternatively, indirect-feedback attenuators might be used.

1.6 Discussion

In the first section of this chapter we formulated some criteria for the optimum adaptation of amplifiers to signal sources and loads. For this purpose we made use of the concept ‘best reproducing relation’ and ‘transmittance’, to emphasise that the required amplifier properties for a given source and load follow from information-transfer aspects. In order to realise accurate and linear transfer of information, the application of negative feedback around an active amplifier part is imperative. For the impedance adaptation to sources and loads, either very low, very high or accurate and linear input and output impedances are necessary. Nine different types of transmittances which can match all types of sources and loads can be distinguished.

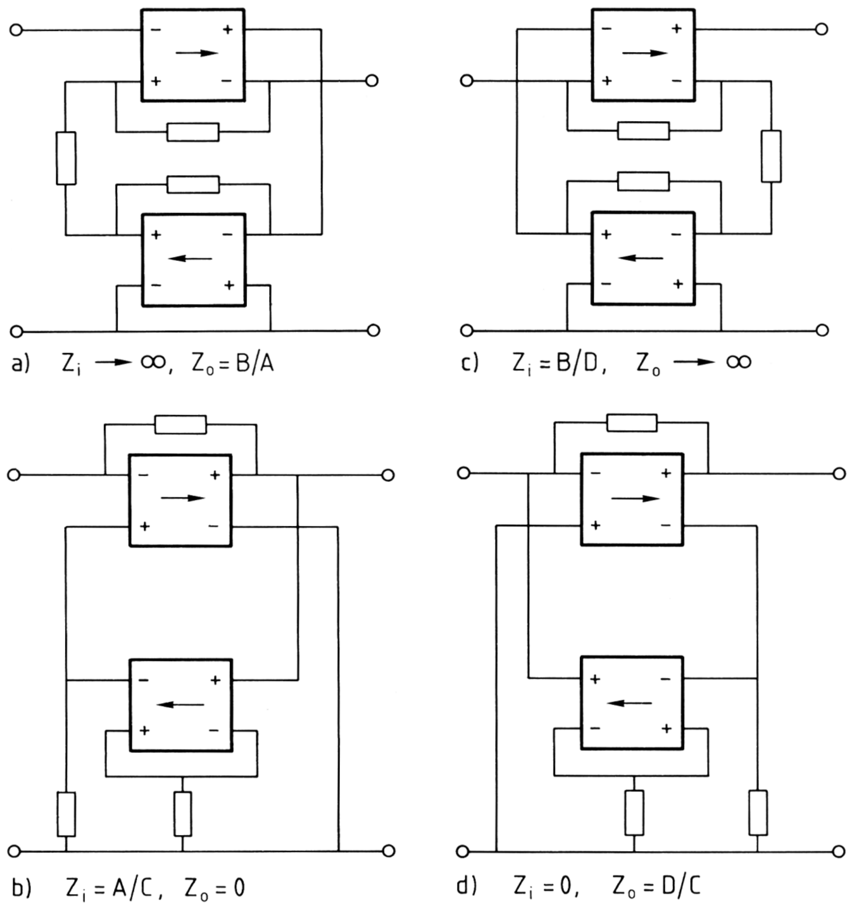


Figure 1.33 Basic amplifier configurations with active feedback loops in addition to passive loops.

A first introductory, largely theoretical approach to the realisation problem made use of ideal feedback two-ports: the ideal transformer and the ideal gyrator. The ideality of these two-ports refers to the absence of series and shunt impedances in the amplifier signal path, which would deteriorate signal-to-noise ratio and efficiency. 'Brute force' terminations are pernicious, and therefore, the proper types of feedback at input and output are needed.

It was shown that 16 basically different configurations exist, one having no feedback loops (the nullor), four with a single loop, six with two loops, four with three loops and one with four loops. Each feedback loop determines one transmission or transfer parameter nearly independently, the reason why a description of the amplifier transfer properties in terms of these parameters is preferred. All desired transmittances can be realised with these basic configurations. The configurations with three feedback loops have no obvious applications.

More practical configurations are found when transformers and impedances are admitted as elements of the feedback networks. This class of feedback configurations, presented in this chapter, is believed to be new. Some of the configurations have been incidentally proposed before, but a systematic classification such as given here has not been found elsewhere. Most configurations with passive impedances, except transformers, in the feedback network are well known, though they have not been classified systematically before either.

Some transmittances cannot be obtained with this class of amplifiers because one of the transfer parameters must have the opposite sign. The application of active feedback offers the possibility to realise the desired additional amplifier types. Two fundamentally different active-feedback methods can be distinguished. The first has been called an indirect-feedback technique because both sensing and comparing are not related to the actual load and source quantities. Accurate device matching is required for this technique, and therefore it is suitable for applications in integrated circuits.

The second is called an active-feedback technique. Sensing and comparing are direct in this case, where one of the conventional passive-feedback configurations or an indirect-feedback configuration is employed in the feedback network.

The basic amplifier configurations presented in this first chapter provide the designer with the possibility to select the right amplifier type for his specific design problem.

Unfortunately, the feedback network generally degrades noise performance as well as power efficiency to some extent. In the subsequent chapters we will address the question of how far these quality aspects are deteriorated. Furthermore, we will deal in detail with the design of the active part of the amplifier.

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