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VGA With Voltage Feedback, Constant Bandwidth and a High Degree of Independence Between Amplifier and Gain Controller

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ABSTRACT The following paper is devoted to an amplifier with gain control (Variable Gain Amplifier). The role of amplification and gain control is addressed separately from the very beginning, to achieve the greatest possible independence of the required parameters both for the amplifier itself (gain size, bandwidth, noise figure) and for the gain controller (energy consumption control, distortion). This feature is very convenient and is not common with other VGA amplifiers. The core of the amplifier consists of an operational amplifier with voltage feedback (VFB) in an inverting circuit, to the feedback of which a gain regulator executed with a diode bridge is connected in parallel. The analysis of the simplified circuit model shows that it is possible to obtain a variable-gain amplifier that maintains its bandwidth. The control of the gain is continuous and only one element, the variable resistance, is sufficient for its control. Continuous control with one element preserving the bandwidth of an operational amplifier with VFB is unique and the authors are unaware of any other similar solution. The proposed VGA amplifier also excels in low noise and negligible distortion. If maximum gain is required, the diode bridge is disconnected and does not contribute to the resulting noise of the VGA amplifier. Because the diode bridge is connected between the output of the operational amplifier and its inverting input, and because the inverting input is current controlled, the diode bridge processes only a portion of the output voltage of the amplifier, thus preventing more noticeable distortion. The distortion is further suppressed due to the simultaneous action of four diodes in the bridge, whose characteristics are reciprocal for the signal path. The presented construction is relatively simple and can be a topic not only for other microelectronic structures, but also for applications from discrete components. The achieved parameters of the presented connection are as follows: gain change 0 to 20 dB, constant bandwidth 7 MHz, distance III. harmonics min. 30 dB at an output voltage of 0.5 V, maximum power spectral density at the output $25 \cdot 10^{-15} \text{ V}^2\text{Hz}^{-1}$.

INDEX TERMS Variable gain amplifier, VGA, voltage feedback, VFB, diode bridge, low distortion, low noise, constant bandwidth amplifier, operational amplifier, automatic gain control, AGC.

I. STATE OF THE ART

Variable gain amplifiers (VGA) can either be stand-alone functional units without feedback, or they are part of automatic gain control and operate in a feedback loop. Automatic gain control circuits ensure a constant level of the output signal even despite fluctuating levels of the input signal. The need to stabilize the signal level can be encountered

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in an inexhaustible number of applications, whether it be computer technology (e.g. reading channels of disk drives), measuring technology (e.g. output circuits of signal generators), medical technology (e.g. input circuits of ultrasound sensors) or telecommunication technology (input circuits of mobile receivers, where the signal level changes with position, or input circuits of fixed receivers, where the signal level changes with a change of transmission channel parameters). It is not possible to use a constant gain value anywhere the input signal level fluctuates. With a strong signal, the level

would be overexcited and the dynamic parameters would deteriorate, further resulting in distortion, and a decrease in the signal-to-noise ratio. Conversely, in the case of a weak signal, the threshold level of the discriminating circuit would not be provided.

Based on papers published to date, variable gain amplifiers can be divided into the following groups:

The first group consists of amplifiers that, by changing the emitter current of the bipolar transistor, control the magnitude of its transconductance and therefore even the gain of the whole stage. Circuit-wise, this can be a single-transistor amplifier stage, a differential pair of transistors, or more complex units, such as a transconductance amplifier (OTA), a current conveyor (CC), a current feedback amplifier (CFA), a Gilbert cell, and others.

An example of a simple controlled amplifier stage connected as a cascode can be found in [1], and a high-frequency adaptation can be found in [2], which arranges several cascode stages in parallel to form a distributed amplifier. [3]–[5]. Whether it is a simple stage or a differential pair, their disadvantage is the relatively small range of input voltages, which can still be processed linearly. This problem can be eliminated by adding emitter resistors to each transistor, however, this comes at the cost of a decrease in differential stage gain. The mentioned modification can be seen in paper [6], in which the controlled differential stage is shifted all the way to after the preamplifier. Another modification is the multiple use of differential stages operating under a common load, the total gain of which is controlled by changing the bias of these stages [7]. The range of input voltages that can be processed linearly increases with decreasing gain.

A transconductance amplifier (OTA) is built on a differential pair of transistors, the output current of which is given by the product of the input differential voltage and the current-controlled transconductance. The range of differential voltage in which the OTA can operate linearly is again not usually greater than 10 mV. If the input differential input voltage is greater, circuit modifications are required, for example, of the linearizing diodes. OTAs were also deployed in CMOS technology. Examples are [8], [9] and [10], which maintain bandwidth even despite changing the gain.

Slightly better properties than traditional transconductance OTA amplifiers, such as lower power consumption and higher bandwidth, are offered by current conveyors (CC). An example is [11], which uses two current conveyors, with the resulting gain given by the ratio of their quiescent currents.

Controlled transconductance is the basis of even more complex units, such as current feedback amplifiers (CFA), whose basic property is a constant bandwidth independent of amplification.

The differential connection of current feedback amplifiers (CFA) is the scope of [12]. Here the gain is given not by the absolute value of the transconductances of the first and second degree, but by their ratio. A gain accuracy of 0.5% is achieved here without the need for trimming. The effects of

temperature and aging are also excluded. The structure is part of the Mixed Signal ASIC in VLSI design.

The differential stage is also part of the Gilbert cell. This cell can already multiply the voltages of both polarities and work in all four quadrants. It also has a larger range of voltage levels that it can process linearly. An example could be [13], or potentially [14], which also maintains bandwidth despite the change in gain.

The second group consists of fixed gain amplifiers preceded by a switchable attenuator. However, the resistive attenuator adds noise at the input of the amplifier. Some papers, therefore, try to circumvent this by introducing a capacitive attenuator [15]. A subset can then be operational voltage feedback amplifiers (VFA), which control the gain by selecting the appropriate tap of the feedback network. These are usually digital controls, and the gain can only be adjusted in discrete steps. They may [16], [17] or may not [18], [19] preserve frequency bandwidth. An exception is the interpolating VGA, again proposed by B. Gilbert [20]. In this amplifier, the gain can be adjusted continuously.

The third group consists of amplifiers that use a FET transistor as a controlled resistor (triode region) to change the gain. The first example can be [21]. Even if it is built on current conveyors, the gain control is left to a voltage divider, which consists of four controlled active resistors - MOS transistors in the triode region. The bandwidth of this amplifier is independent of the set gain. Another paper [22] achieves a change in gain by lesser or greater degeneration of a multiple differential amplifier using active resistors - MOS transistors in their emitters. The paper [23] describes a controlled cascode amplifier with MOS transistors, in which the innovation concerns the cascode transistor. If its bias voltage is not fixed but adjustable, the transistor can work under certain conditions either as a controlled resistor (triode region) in the CS circuit, or in the saturation region in the CG circuit, which is a classical cascode, and thus adjust the gain. Active feedback is introduced to achieve high bandwidth and large gain changes. An amplifier with gain control in the range of 16 dB for ultrasonic applications in the 10 kHz to 2 MHz band is described in [24]. Its connection is executed using discrete components and consists of an operational amplifier, a resistive network and two J-FET transistors that operate in the triode region. J-FET transistors are connected in both the direct and the feedback path of the operational amplifier, whereas the change in gain is achieved by means of a transistor in the feedback path.

The fourth group consists of amplifiers with PIN diodes which, due to the physical principle of PIN diodes, are predetermined for high frequencies. The paper [25] describes the design of an amplifier with a medium frequency of 1 GHz, which uses a PIN diode to control the transmission of feedback and thereby also the size of the gain. Dual gain management is the subject of [26]. In addition to the PIN diode in the feedback, the quiescent current of the output bipolar transistor in the CC connection is also controlled. This amplifier operates in the 3 to 10 GHz band.

There are, of course, numerous perspectives for division, and the overview given is just one of many. Some papers may fit into several groups at the same time. It is also not in the abilities of the authors to provide a definitive list of all solutions, and this is of course not the aim of this paper.

II. MOTIVATION FOR A NEW SOLUTION

The above overview of solutions points out several important facts:

a) Most of the gain-controlled amplifiers mentioned above describe advanced microelectronic structures that are capable of functioning and achieve their advantages only when integrated on a single chip and thermally coupled.

Due to the costs of developing integrated circuits, only those structures are integrated which will find mass application, for example in communications, computer technology, etc., and which will ensure a return on investment. However, the use of VGAs in areas of technology where the available integrated VGA structures may not be optimal is increasing. One such area, for example, is that of optoelectronics. A typical example is an electro-optical measuring system using the Pockel effect [27], in which a VGA is used to eliminate a systematic error and with it the noise of a laser source. Another example may be fiber optic mechanical stress sensors using Bragg gratings [28]. Here, a VGA is used to eliminate the effect of ambient temperature. Other areas include medical technology and measuring technology in general. It is worth mentioning the software loop of automatic gain control using digital potentiometers [29], which is part of an ECG measuring system without a reference electrode. Other VGA applications can be found in research and development of new products, whether in piece laboratory verification or mass testing in series production. As a general example, we can mention [30], which describes an adaptive converter of light intensity to an electrical signal, the core of which is a controlled transimpedance amplifier. Another incentive for a new VGA solution might also be the almost too low supply voltage of modern integrated technologies. We also must not neglect ad-hoc solutions, which are based, for example, on the available component base, or situations where quick solutions are required.

The aim of the paper therefore, was to find a VGA solution that can serve not only as another idea for a microelectronic structure, but which for its simplicity will not be a problem to execute using discrete components available on the market.

b) A relatively small percentage of papers is devoted to operational amplifiers with voltage feedback. Given the growing bandwidth requirements of VGA amplifiers, this is understandable, since the dynamic parameters of VFAs (bandwidth, ramp rate) can no longer compete with the dynamic parameters of CFAs. On the other hand, VFA operational amplifiers are much more widespread, and the portfolio of operational voltage feedback amplifiers, whose GBW normally exceeds 300 MHz, has also grown in the last decade. Most work dealing with VFA operational amplifiers offers only discrete gain control, which is compatible

with microprocessor control, but not with analog control of autonomous AGC systems.

The authors therefore decided to look for new ways of continuous gain control in operational amplifiers with voltage feedback (VFA).

c) For VGA amplifiers, a wide range of parameters are evaluated, namely, for example, the range of gain change, linearity of gain depending on the control voltage or current (not so critical in AGC systems), noise figure, input impedance, signal distortion (IP3), power consumption, control method (continuous, discrete), etc. Other important parameters also include the ability to maintain bandwidth despite a change in gain (this feature is very valuable in measurement technology and laboratory practice, especially when measuring noise parameters of electronic components).

Most of the integrated structures mentioned above are closed solutions with certain parameters optimized for one or another particular application. These parameters usually cannot be influenced. However, many designers would welcome a modular system, a universal VGA amplifier block, which could be cascaded and, at least in terms of gain, adapted to the given application. To make cascading possible, the VGA should be able to linearly process even a larger input signal, as well as to reduce the gain to 0 dB. This is also required if the VGA is arrayed after a fixed gain amplifier (e.g., for low noise circuits, which the first stage usually solves as fixed and noise optimized). If the VGA amplifier is counted on being implemented immediately in the first amplifier stage, it should have the lowest possible noise figure (the papers cited above have, with a few exceptions, relatively high noise figures). Furthermore, the designer would appreciate the possibility of setting the level of distortion (IP3) independently of other parameters, or potentially good DC parameters of the entire amplifier, such as low offset and drift.

Critically speaking, it should be noted that the given list of requirements is too large for a single circuit structure. It would therefore be good to divide these tasks between the amplifier and the gain controller, i.e., to ensure a certain degree of independence of the amplifier and the gain controller – ideally, to design the amplifier and the gain controller separately.

The idea of addressing the amplifier and gain regulator separately, as two independent units, is not new. This idea was first mentioned in [31], which complements a standard fixed-gain VFA amplifier with a gain controller. This is added as controlled feedback from the output of the operational amplifier to its inverting input (sum point) in parallel with the fixed feedback. The core of the gain regulator is a voltage controlled current source. Although this is an interesting solution, it has the following shortcomings:

- a) The amplifier is relatively narrowband
- b) The amplitude frequency response characteristic shows a large ripple (up to 7 dB), which in addition fluctuates with the size of the set gain
- c) The amplifier does not maintain the gain bandwidth with a change in gain

d) The controlled feedback contains another active element that may lead to potential instability

e) It cannot be adapted for different levels of input (output) signal and constant distortion

f) The paper does not address the size of the R_{FB} resistor, which is very important in terms of setting the operating point of the operational amplifier, while its size may affect the size and price of the chip.

Therefore, another way is sought to best solve VGA as two independent tasks: a) amplifier, b) gain control. The amplifier could be selected on the basis of DC, dynamic and noise parameters, and the design of gain control then on the basis of maximum distortion at a specific level of the input signal and the set gain. It would also be desirable for the amplifier to not affect the operation of the gain controller and for the gain controller to not degrade the DC, dynamic, or noise parameters of the amplifier.

The paper further deals with a voltage-feedback VGA amplifier, which offers a constant bandwidth and a large degree of independence of the amplifier and gain controller. The basic features of the presented connection are described in the chapter entitled A New VGA Concept. In the chapter entitled Basic VGA Circuit Design, a specific design of amplifier and gain controller circuits is presented. The outputs from the measurement of frequency response characteristic, distortion and noise of the designed amplifier are presented in the chapter Parameters of the Basic VGA Circuit Design. The chapter Constant Bandwidth VGA Wiring deals with circuit modifications leading to constant bandwidth and simultaneously presents updated frequency response characteristics. The chapter VGA Gain Control deals with the functional dependences of the control voltage and the set gain. How to achieve good DC parameters of the amplifier as a whole is addressed in the chapter VGA Offset Elimination.

III. A NEW VGA CONCEPT

As follows from the previous analysis, an amplifier with a continuous change of gain is built in such a way as to ensure the greatest possible degree of independence of the amplifier and the gain controller. The amplification is provided by an operational amplifier with voltage feedback in the inverting wiring. This connection is advantageous because the potential of the inverting and non-inverting terminals is constant during the operation of the operational amplifier. It is then possible to lean the control voltage of the gain controller against the node with a constant potential, or to connect the gain controller to an operational amplifier without alternating coupling and limiting the lower transmitted frequencies.

A simple resistor network composed of resistors R_{amp1} and R_{amp2} , Fig. 1(a) and Fig. 1(b), which are connected to the inverting terminal of the operational amplifier, is sufficient for the operation of the operational amplifier with an inverting connection. The gain of the operational amplifier in the inverting connection is then given by:

$$A = \frac{u_{OUT}}{u_{IN}} = -\frac{R_{amp2}}{R_{amp1}} \quad (1)$$

Since only an operational amplifier and two resistors are needed for this amplification, one can expect relatively low noise in such an amplifier and optimal conditions for processing very small signals. Further, it is appropriate here to define the feedback network transfer β , where for an inverting operational amplifier circuit:

$$\beta = \frac{u^-}{u_{OUT}} = \frac{R_{amp1}}{R_{amp1} + R_{amp2}} = \frac{1}{1 + \frac{R_{amp2}}{R_{amp1}}} = \frac{1}{1 + |A|} \quad (2)$$

And further, the upper limit frequency f_{CL} of the closed loop of the operational amplifier and feedback network loop:

$$f_{CL} = \beta f_T \text{ [Hz]} \quad (3)$$

Meanwhile, u^- is the voltage of the inverting terminal and f_T the upper limit frequency of the operational amplifier.

We then let the amplification expressed by relation (1) be the maximum possible and let potential regulation of the gain address only the decrease of the gain. This will be achievable either by increasing the resistance of the resistor R_{amp1} , Fig. 1(a), or by decreasing the resistance of the resistor R_{amp2} , Fig. 1(b). The first option is less advantageous because it arranges the gain controller in series with the resistor R_{amp1} , Fig. 1(a). If no mechanical contacts are used that would completely disconnect the gain controller in an inactive state, the transition to the inactive state will be achieved only by switching the gain controller to low impedance. In this state, however, the noise of the gain controller continues to act, and the noise of the operational amplifier and its network are added together, increasing the total noise at the output of the VGA amplifier. It will therefore be more advantageous to arrange the gain controller in parallel with the resistor R_{amp2} , Fig. 1(b). Thanks to the parallel connection, the inactive state of the gain controller will be achieved by its transition to high impedance. Any noise of the gain controller then disappears at the very small impedance of the summing node of the operational amplifier, or at the very small impedance of the output of the operational amplifier, because these small impedances together with the large internal resistance of the gain controller create a resistance divider with very low transmission.

Further considerations will therefore focus on the circuit of Fig. 1(b). A dipole is sought which will be able to provide a high impedance between its terminals if necessary, or vice versa, the lowest possible impedance. The solution is relatively simple and can be found among the sampling circuits. The requirement is well met by the sampling diode bridge, Fig. 2. However, its function will not be limited only to the ON or OFF states, but it will also serve between these limit states. This can be achieved thanks to the continuous VA characteristic of the diode.

The current through the silicon diode can be approximated by an exponential function:

$$I = I_0 \left(e^{\frac{U}{2U_T}} - 1 \right) \text{ [A]} \quad (4)$$

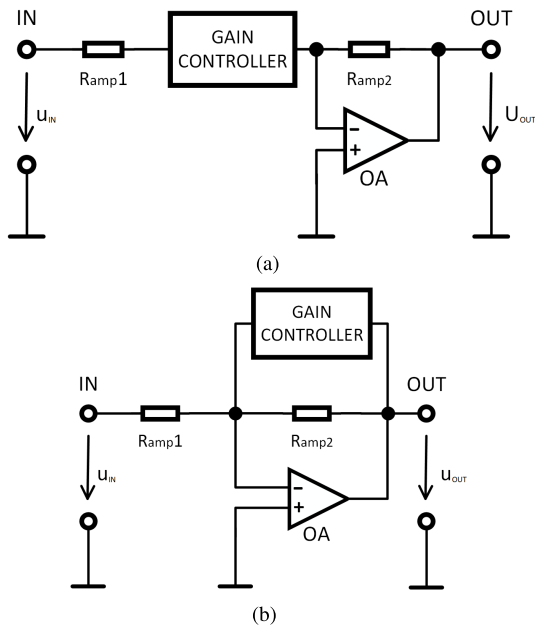


FIGURE 1. Wiring possibilities of an inverting operational amplifier with a gain controller, a) in series with a resistor R_{amp1} , b) in parallel with a resistor R_{amp2} . The gain regulator includes variable resistance and control circuits.

The voltage on the silicon diode can be found by logarithmizing the relationship (4):

$$U = 2U_T \ln \left(\frac{I}{I_0} + 1 \right) \text{ [V]} \quad (5)$$

In relation (4) and (5), U represents the terminal voltage of the diode, I the current through the diode, I_0 the saturation current of the diode in the closing direction and U_T the temperature voltage, which is calculated according to the relation:

$$U_T = \frac{kT}{q} \text{ [V]} \quad (6)$$

where as k is the Boltzmann constant ($k = 1.38064852 \cdot 10^{-23} \text{ m}^2 \text{ kg s}^{-2} \text{ K}^{-1}$), T is the thermodynamic ambient temperature in Kelvin and q is the electron charge ($q = 1.60217662 \cdot 10^{-19} \text{ C}$). It is more advantageous to derive the dynamic resistance of a silicon diode from relation (5) by its derivation according to the current:

$$r_d = \frac{dU}{dI} = \frac{2U_T}{I + I_0} \text{ [\Omega]} \quad (7)$$

For a standard diode, e.g., 1N4148 (Vishay, Malvern, PA, USA, 2017), the saturation current is usually $I_0 \approx 25 \text{ nA}$ and for room temperature $T = 300 \text{ K}$ the temperature voltage is $U_T = 26 \text{ mV}$. The magnitude of the dynamic resistance of a closed diode is found by substituting into relation (7) for a current $I = 0 \text{ A}$, which is $r_{dmax} = 2.08 \text{ M}\Omega$. Conversely, the magnitude of the dynamic resistance of an open diode is found by substituting into relation (7) for a specific, nonzero current I , for example, $I = 250 \mu\text{A}$, which is $r_{dmin} = 208 \Omega$.

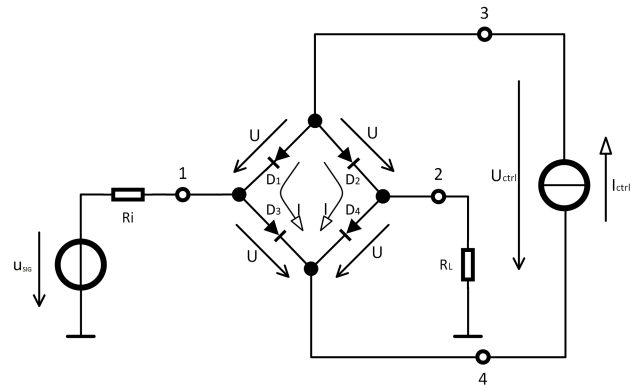


FIGURE 2. Diode bridge in the function of variable resistance of the gain controller.

Clearly, the magnitude of the saturation current I_0 , has the greatest influence on the dynamic resistance of the closed diode, while the current I in the forward direction has the greatest influence on the dynamic resistance of the open diode. It is obvious that for diode I currents between 0 and $250 \mu\text{A}$ the dynamic resistance of the diode r_d will be between $2.08 \text{ M}\Omega$ and 208Ω .

In terms of signal path, the diode bridge appears as a dipole, but in fact it has a total of four poles. The poles, or also terminals 1 and 2, are signal terminals whose mutual impedance is controlled by the current I_{ctrl} flowing between control terminals 3 and 4. If the diode bridge is made of identical diodes (preferably on one chip with thermal coupling) and if the signal and control paths are properly separated, the signal and control currents are independent of each other and do not affect each other. For further considerations, let terminals 3, 4 be connected to a current source I_{ctrl} with infinitely large impedance.

The diode bridge connects the source and the load, i.e., signal terminals 1 and 2, using a series-parallel combination of dynamic resistors r_d of diodes D_1 to D_4 . The magnitude of the dynamic resistance is determined by the position of the operating point on the VA characteristic of each of the diodes according to the magnitude of the flowing direct current I . With identical diodes D_1 to D_4 and identical currents I in branches D_1 - D_3 and D_2 - D_4 , the same dynamic resistances can be expected for the diodes r_d . The resulting dynamic resistance will be given by:

$$r_{dfinal} = \frac{(r_{d1} + r_{d2})(r_{d3} + r_{d4})}{r_{d1} + r_{d2} + r_{d3} + r_{d4}} = |r_d = r_{d1} = r_{d2} = r_{d3} = r_{d4}| = r_d \text{ [\Omega]} \quad (8)$$

The total dynamic resistance r_{dfinal} between signal terminals 1 and 2 corresponds to the dynamic resistance r_d of one diode. The limits of dynamic resistance r_{dfinal} of the whole bridge will therefore be identical with the limits of dynamic resistance r_d of one diode, which were the values $r_{dfinalmin} = r_{dmin} = 208 \Omega$ and $r_{dfinalmax} = r_{dmax} = 2.08 \text{ M}\Omega$. Furthermore, the diode bridge

according to Fig. 2 is reciprocal, so it does not matter which of terminals 1 or 2 will be connected to the source or to the load.

Thus, let the diode bridge described above be connected in parallel to the resistor R_{amp2} , i.e., terminal 1 to the summing node of the operational amplifier OA and terminal 2 to the output of the operational amplifier OA. If maximum VGA gain is required, the diode bridge is disconnected and the power input of the bridge control circuits is zero. The control current I_{ctrl} is zero and the control voltage of the bridge U_{ctrl} , is zero, as is then the current I and the voltage of the diodes U . Under these conditions, the internal resistance of the bridge will be the greatest, close to $r_{dfinalmax}$. As mentioned earlier, all the noise sources of the bridge diodes and the noise sources of the control circuits will be divided towards the operational amplifier OA in the ratio r_{OA}/r_{dmax} .

Whether it is the resistance r_{OA} of the input inverting terminal or the output terminal of the operational amplifier, the resistance r_{OA} will be close to zero and the transmission to the operational amplifier from the bridge side will also be close to zero. This is evidenced by the measurement of the noise voltage at the VGA output, which remained unchanged for the “stand-alone amplifier” and “diode bridge deactivated” configurations. Details can be found in the section entitled Parameters of the Basic VGA Circuit Design. Because the bridge processes small signal voltages and because its capacity does not exceed a few picofarads, it is not necessary to provide a disconnected bridge with a negative bias at diodes D_1 to D_4 , or at terminals 3 and 4.

If, on the other hand, minimal VGA gain is required, a diode bridge is connected. If the current passing through diodes D_1 to D_4 is equal to $I = 250 \mu A$, then the current I_{ctrl} between control terminals 3 and 4 is equal to $I_{ctrl} = 2I = 0.5 \text{ mA}$. If the voltage on the diode is $U = 0.479 \text{ V}$ (according to the approximation above), then the bridge voltage between control terminals 3 and 4 is equal to $U_{ctrl} = 2U = 0.958 \text{ V}$. In this state, the power for controlling the diode bridge is highest and is found according to:

$$P_{ctrl} = U_{ctrl}I_{ctrl} = 2U2I = 4UI \text{ [W]} \quad (9)$$

After substituting for $I_{ctrl} = 0.5 \text{ mA}$ and $U_{ctrl} = 0.958 \text{ V}$, $P_{ctrl} = 479 \mu W$. Although this value is not in any way large, the need for a larger control current I_{ctrl} , which is 0.5 mA for the above case, may cause a problem. The total power input of the bridge control can of course be greater than P_{ctrl} and is affected by other control circuits. The internal resistance of the bridge will be the smallest under these conditions, close to $r_{dfinalmin}$.

Furthermore, the issue of distortion needs to be addressed. Although the solution of Fig. 1(a) has been discarded for unfavorable noise conditions, it offers very little distortion. This is due to the very small voltages and currents that act between the input terminal IN and the summing node (inverting terminal) of the operational amplifier. This may no longer be true for the chosen solution from Fig. 1(b). If an amplitude of, for example, 1 V is required as the output voltage of the operational amplifier, this amplitude will act

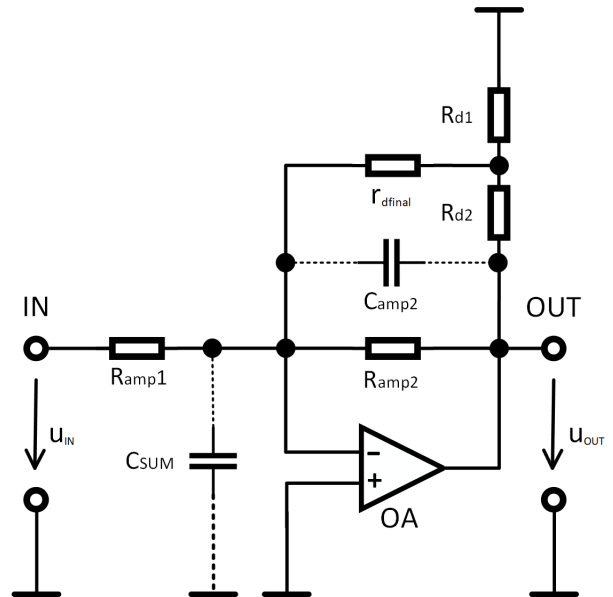


FIGURE 3. Substitute model of the designed VGA with a diode bridge.

between the output terminal OUT and the summing node (inverting terminal) of the operational amplifier, and also between terminals 1 and 2 of the diode bridge. However, the transmission of a 1 V voltage fluctuation across the bridge will cause noticeable distortion. However, there is a solution: Because the summing node of the operational amplifier is current excited, a certain signal current level can be achieved for a larger value of r_{dfinal} resistance, with a greater voltage amplitude at the input of the bridge, or for a smaller value of r_{dfinal} resistance with a smaller voltage amplitude at the input of the bridge. Therefore, if the smaller r_{dfinal} values are not detrimental, the output voltage of the operational amplifier towards the bridge can be divided by a resistance divider R_{d1} and R_{d2} . The resistive divider maintains a large voltage amplitude at the output of the operational amplifier and ensures a small voltage amplitude at the input of the diode bridge and thereby also a small distortion of the processed signal, Fig. 3. The feedback network of the operational amplifier will therefore consist of the basic divider of R_{amp2} , R_{amp1} , and a controlled T-element, R_{d2} , R_{d1} and r_{dfinal} , which will be connected in parallel to the resistor R_{amp2} . The resistors R_{d2} , R_{d1} of the auxiliary divider will be fixed, while the value of the resistance r_{dfinal} will vary depending on the control current I_{ctrl} of the diode bridge.

Since the smallest possible distortion and hence the largest possible separation ratio $R_{d1}/(R_{d1} + R_{d2})$, is required, it will be necessary to determine the interdependence of the separation ratio $R_{d1}/(R_{d1} + R_{d2})$ and the dynamic resistance of the bridge r_{dfinal} and how it is affected. For wiring according to Fig. 3, the following applies:

$$\frac{u_{IN}}{R_{amp1}} + \frac{u_{OUT}}{R_{amp2}} + \frac{u_{OUT} \frac{R_{d1}}{R_{d2} + R_{d1}}}{r_{dfinal} + \frac{R_{d1}R_{d2}}{R_{d1} + R_{d2}}} = 0 \quad (10)$$

At minimum amplification $A_{min} = -1$, the magnitude of the input and output voltage will be the same $-u_{OUT} = u_{IN}$

and the dynamic resistance of the bridge will be the smallest, $r_{dfinal} = r_{dfinalmin}$:

$$-\frac{1}{R_{amp1}} + \frac{1}{R_{amp2}} + \frac{\frac{R_{d1}}{R_{d2}+R_{d1}}}{r_{dfinalmin} + \frac{R_{d1}R_{d2}}{R_{d1}+R_{d2}}} = 0 \quad (11)$$

If R_{amp1} a R_{amp2} were calculated for maximum amplification, $|A| \gg 1$, for example $|A| = 10$, it is possible to overlook the term $1/R_{amp2}$ in (11) and re-write the equation:

$$\frac{r_{dfinalmin} + \frac{R_{d1}R_{d2}}{R_{d1}+R_{d2}}}{R_{amp1}} \doteq \frac{R_{d1}}{R_{d2} + R_{d1}} \quad (12)$$

The maximum possible division ratio R_{d2}/R_{d1} is thus approximately determined by the smallest dynamic resistance of the diode bridge $r_{dfinalmin}$, the internal resistance of the divider $R_{d1}||R_{d2}$ and the input resistance of the amplifier R_{amp1} . The maximum division ratio R_{d2}/R_{d1} , requires the lowest possible dynamic resistance $r_{dfinalmin}$, the smallest possible internal resistance of the divider $R_{d1}||R_{d2}$ ($R_{d1}||R_{d2} < r_{dfinalmin}$), and vice versa, the largest possible resistance R_{amp1} . However, the resistance of R_{amp1} cannot be increased unchecked, because its value together with the capacity of the sum node of the operational amplifier, C_{sum} , determines the final dynamic parameters of the VGA amplifier:

$$f_{pool} \doteq \frac{1}{2\pi R_{amp1} C_{sum}} [Hz] \quad (13)$$

Though the capacity of the sum node C_{sum} can be compensated for by the capacity of C_{amp2} , connected in parallel to the resistor, R_{amp2} ,

$$f_{pool} \doteq \frac{1}{2\pi R_{amp2} C_{sum2}} [Hz] \quad (14)$$

$$f_{null} \geq f_{pool} \quad (15)$$

Nevertheless, from the limit frequency, f_{null} , the maximum possible gain is the given not by the resistance ratios R_{amp2} and R_{amp1} , but by the ratio of the capacities of C_{sum} and C_{amp2} . In addition, the variable capacity of the bridge depending on the current I_{ctrl} is included in the capacity of C_{amp2} . It is therefore better to keep the resistance values of the feedback network as small as possible and the pole frequency, f_{pool} , as large as possible.

If the minimum value of the dynamic resistance of the bridge $r_{dfinalmin} = 208 \Omega$, the internal resistance of the divider $R_{d1}||R_{d2} \approx 100 \Omega$ and the maximum value of the resistance, for example $R_{amp1} = 2.2 k\Omega$, are considered, the transmission of the divider R_{d1} , R_{d2} according to relation (12) is around 0.1, i.e., the output voltage of the operational amplifier can be divided towards the bridge approximately 10x. Proportionate to this value, the distortion of the output signal also decreases. If it were possible to increase the control current of the diode bridge up to the value $I_{ctrl} = 10 \text{ mA}$ and work with the current through the diode $I = 5 \text{ mA}$, the dynamic resistance of the bridge would decrease all the way to $r_{dfinalmin} = 10.4 \Omega$. Upon reduction

of the internal resistance of the divider to $R_{d1}||R_{d2} \approx 10 \Omega$ while maintaining the resistor $R_{amp1} = 2.2 k\Omega$, then, according to relation (12), the transmission of the divider R_{d1} , R_{d2} comes out at up to around a value of 0.01, i.e., the output voltage of the operational amplifier can be divided towards the bridge approximately 100x and distortion at the output of the operational amplifier can be eliminated by that much more. When changing the division ratio of R_{d1} , R_{d2} , it is not necessary to change the value of the resistance R_{d2} , which determines the load on the output of the operational amplifier. In both cases, $R_{d2} = 1 k\Omega$ is used, and the output of the operational amplifier is thus not overloaded. The R_{d2} value must also be monitored to achieve the smallest required transmission, A_{min} . As will be shown below, if $A_{min} = -1$ is required, the resistance R_{d2} should be approximately equal to the resistance R_{amp1} . It can be seen that the design of the gain regulator and of the amplifier itself are not mutually dependent in any significant way. Their only connecting link remains the resistor R_{amp1} .

Based on the previous considerations, it is clear that the VGA wiring according to Fig. 1(b) held up both from the perspective of noise (diode bridge disconnected) and of distortion (diode bridge connected). It will therefore be appropriate to derive and investigate its voltage transfer. This can be obtained by modifying the relationship (10):

$$A = -\frac{R_{amp2}}{R_{amp1}} \cdot \frac{r_{dfinal} \frac{R_{d1}+R_{d2}}{R_{d1}} + R_{d2}}{r_{dfinal} \frac{R_{d1}+R_{d2}}{R_{d1}} + R_{d2} + R_{amp2}} [V] \quad (16)$$

It is clear that if the diode bridge is disconnected, $r_{dfinal} \rightarrow \infty \Omega$, the relationship (16) shifts into its basic form (1):

$$\begin{aligned} \lim_{r_{dfinal} \rightarrow \infty} A &= -\frac{R_{amp2}}{R_{amp1}} \cdot \frac{\frac{r_{dfinal}}{r_{dfinal}} \frac{r_{d1}+r_{d2}}{r_{d1}} + \frac{r_{d2}}{r_{dfinal}}}{\frac{r_{dfinal}}{r_{dfinal}} \frac{r_{d1}+r_{d2}}{r_{d1}} + \frac{r_{d2}}{r_{dfinal}} + \frac{r_{amp2}}{r_{dfinal}}} \\ &= -\frac{R_{amp2}}{R_{amp1}} \cdot \frac{\frac{R_{d1}+r_{d2}}{R_{d1}}}{\frac{R_{d1}+r_{d2}}{R_{d1}}} = -\frac{R_{amp2}}{R_{amp1}} \end{aligned} \quad (17)$$

If, on the other hand, the diode bridge is connected and $r_{dfinal} \rightarrow 0 \Omega$, the expression (16) changes as follows:

$$\begin{aligned} \lim_{r_{dfinal} \rightarrow 0} A &= -\frac{R_{amp2}}{R_{amp1}} \cdot \frac{r_{dfinal} \frac{r_{d1}+r_{d2}}{r_{d1}} + r_{d2}}{r_{dfinal} \frac{r_{d1}+r_{d2}}{r_{d1}} + r_{d2} + r_{amp2}} \\ &= -\frac{R_{amp2}}{R_{amp1}} \cdot \frac{R_{d2}}{R_{d2} + R_{amp2}} = -\frac{R_{amp2}||R_{d2}}{R_{amp1}} \end{aligned} \quad (18)$$

The relationship (18) determines the maximum value of R_{d2} in relation to the minimum required gain, A_{min} . Relationships (18), (13) and (12) determine the basic design rules of a diode bridge VGA amplifier. The relationship (1) is no longer critical and confirms the independence of the operational amplifier and its passive network on its own diode bridge.

If a diode bridge is connected, the feedback of the operational amplifier is extended by a T-cell, which is used from time to time, for example, in current-voltage converters, where it replaces a less available electrometric resistor with

lower value resistors. However, the use of a T-cell has two basic disadvantages, namely an increase in the input voltage offset and noise at the output of the operational amplifier in the R_{d2}/R_{d1} ratio. It will therefore be necessary to verify whether this disadvantage does not also apply to the existing circuit according to Fig. 3. To do this, it is necessary to derive the expression for NG , Noise Gain, which is the inverse value of the feedback network β defined by (2):

$$NG = \frac{1}{\beta} = \dots (\text{see in Appendix IX}) \quad (19)$$

As can be seen from Fig. 3, the output of the operational amplifier is connected to the summing node of the operational amplifier not only via the T-cell R_{d2} , R_{d1} and r_{dfinal} , but also via the resistor R_{amp2} . It would therefore be good to trace how the change of the R_{amp2} value or the removal of the R_{amp2} resistor ($R_{amp2} \rightarrow \infty$) is reflected in the magnitude of the noise gain NG (f). For this purpose, the numerator and denominator of expression (17) are divided by the value R_{amp2} . Expression (19) changes into the following form:

$$NG = \dots (\text{see in Appendix IX}) \quad (20)$$

If the value of R_{amp2} increases, the numerator of relation (20) will change only slightly, because for maximum gain $|A| \gg 1$, usually $R_{amp2} \gg R_{amp1}$, so the terms R_{amp1}/R_{amp2} would not have to be considered in parentheses at all. However, a big change awaits the denominator, since any, even slight, increase in R_{amp2} will be accompanied by a decrease in the size of the denominator and thus an increase in NG (f). The highest noise gain NG (f) therefore belongs to the limit value $R_{amp2} \rightarrow \infty \Omega$:

$$\lim_{R_{amp2} \rightarrow \infty} NG = \left(1 + \frac{R_{d2}}{R_{d1}}\right) \left(1 + \frac{R_{dfinal}}{R_{amp1}}\right) + \frac{R_{d2}}{R_{amp1}} \quad [/] \quad (21)$$

Relationship (21) corresponds to the noise gain of the T-cell itself. The term $(1+R_{d2}/R_{d1})$ then corresponds to the dreaded increase in input voltage offset and noise in the current-voltage converter, but with $R_{amp1} \rightarrow \infty \Omega$ (current excitation at the input). In the discussed inverting amplifier, the situation does not worsen, because the value of R_{amp1} is small, in the order of $k\Omega$. The gain is only divided into two stages, namely the basic term gain $(1+r_{dfinal}/R_{amp1})$ and the additional term gain $(1+R_{d2}/R_{d1})$ mediated by the T-cell. If a resistor, R_{amp2} , is connected, the noise gain further decreases to the value according to the relationship (20).

Thus, the circuit of Fig. 3 does not increase the input voltage offset or the noise of the operational amplifier towards its output. Further analysis of Fig. 3 and relation (19), on the other hand, shows that this is probably the only connection of the operational amplifier that retains the bandwidth despite the change in gain! The stated change of gain, meanwhile, is achieved by means of one element (r_{dfinal}) without any intervention in the rest of the feedback network of the operational amplifier. Proof should be the further adjustment and analysis of the relationship (19). Both the numerator and

denominator are first divided by R_{amp1} and then R_{d1} :

$$NG = \dots (\text{see in Appendix IX}) \quad (22)$$

We let the further transmissions of the dividers R_{amp2} , R_{amp1} and R_{d2} , R_{d1} be the same:

$$D = \frac{R_{d1} + R_{d2}}{R_{d1}} = \frac{R_{amp1} + R_{amp2}}{R_{amp1}} \quad [/] \quad (23)$$

Then, relationship (22) can be rewritten as follows:

$$NG = \frac{r_{dfinal}D^2 + R_{d2}D + R_{amp2}D}{r_{dfinal}D + R_{d2} + R_{amp2}} = \frac{D(r_{dfinal}D + R_{d2} + R_{amp2})}{r_{dfinal}D + R_{d2} + R_{amp2}} = D \quad (24)$$

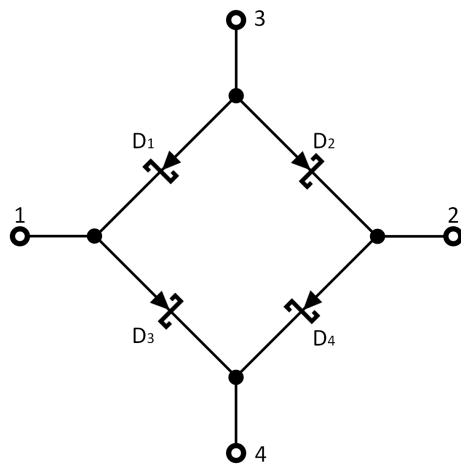
Under these conditions, the noise gain is constant, equal to D and independent of the r_{dfinal} value. Its inverse value $1/NG = \beta$ is also constant, and therefore so too is the resulting bandwidth, $f_{CL} = \beta f_T$. Since the inverse of $1/NG = \beta$ coincides with relationship (2), the resulting bandwidth is the same as for the basic wiring of the inverting operational amplifier.

The feedback circuit of Fig. 3 is actually a resistor bridge. If condition (23) applies, the said bridge on the output side of the operational amplifier, which acts diagonally, is balanced and the gradient on the r_{dfinal} resistance is zero. Therefore, changing the r_{dfinal} resistance does not affect the value of the feedback network β transmission. However, the conditions for the input terminal of the amplifier are different. The input voltage u_{IN} acts in the arm of the bridge, causing its unbalancing and thus the voltage drop to r_{dfinal} . Conversely, the value of r_{dfinal} therefore contributes to the magnitude of the voltage at the output of the operational amplifier.

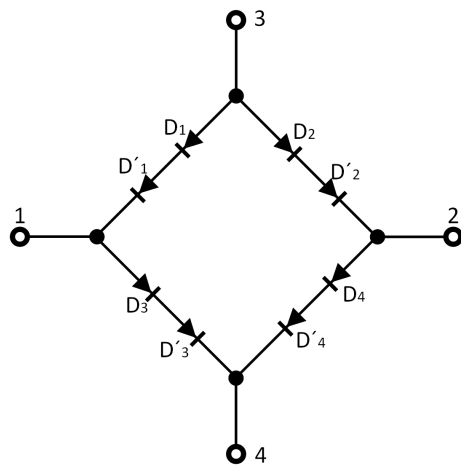
If the resistance r_{dfinal} is created with a two-pole element, e.g. a rheostat, the wiring of Fig. 3 maintains a constant bandwidth. However, if the r_{dfinal} resistor is replaced with a more complex structure, such as a diode bridge, it will depend on the impedance of the bridge control path. In the case that the impedance of the bridge control path is infinitely large, which was, for that matter, the assumption in deriving the relationship (8), a constant bandwidth is maintained. If the control path impedance is small, there is some unbalancing and a constant bandwidth can no longer be expected. For ways to achieve high bridge control path impedance, see the chapter entitled Constant Bandwidth VGA Wiring.

Another advantage of the presented diode bridge is its passive behavior and appliance character. Therefore, there is no need to worry about deterioration in the stability of the operational amplifier, even when used with the bridge in the active state. The disadvantage of the diode bridge is then a certain degree of distortion resulting from the use of nonlinear VA characteristics of the diodes, which are necessary to change the dynamic resistance when changing the operating point and, additionally, the need for four diodes with identical VA characteristics.

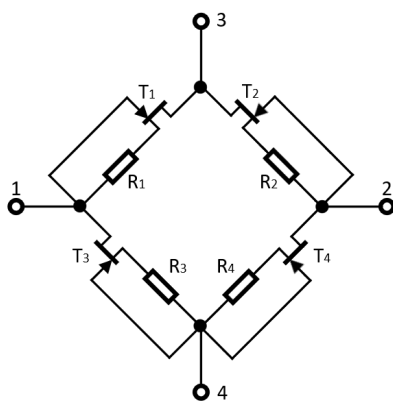
The high degree of independence of the bridge from the amplifier allows both a wide selection of the operational



(a)



(b)



(c)

FIGURE 4. Other circuit solutions of the bridge as a variable resistance of the gain controller. The bridge is fitted with a) Schottky diodes, b) pairs of diodes, c) J-FET transistors.

amplifier itself in terms of frequency, noise and DC parameters, as well as a wide selection of elements with non-linear characteristics for the bridge. In addition to the silicon diodes

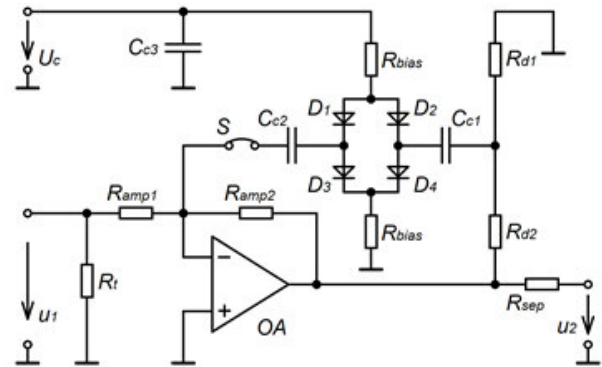


FIGURE 5. Basic VGA circuit design with a diode bridge.

discussed, the bridge can be equipped with, for example, Schottky junction diodes, diode pairs or suitably wired J-FET transistors. These possibilities are shown in Fig. 4. Finally, the mentioned J-FET transistors also reverse the direction of control so that a lower power input will correspond to the minimum VGA gain, and a higher power, then, to the maximum VGA gain. The selection of a suitable, non-linear, element will depend on the allowable distortion and the permissible control current.

IV. BASIC VGA CIRCUIT DESIGN

The amplifier described above was first verified wired according to Fig. 5. In this configuration, the diode bridge is connected to the operational amplifier OA by means of an alternating coupling C_{c1} and C_{c2} . The role of the resistors R_{amp1} , R_{amp2} , R_{d1} , R_{d2} and diodes D_1 to D_4 has already been explained above. The diagram is supplemented with a resistor R_t , which terminates the input coaxial line, and a resistor R_{sep} , which separates the output of the operational amplifier OA from any capacitive load and thereby ensures the unconditional stability of the stage. The R_{bias} resistors connect the control terminals of bridge 3 and 4 to the control voltage U_c and linearize the dependence of the control current I_{ctrl} on the control voltage U_c . A total of two R_{bias} resistors are connected symmetrically to the bridge to ensure the same impedance of terminals 3 and 4 to ground and the same behavior of the bridge for both the positive and negative half-waves of the signal voltage. Jumper S is used for noise investigation. The noise of the operational amplifier OA and the basic passive network R_{amp1} and R_{amp2} is measured with the jumper pulled out and with the jumper inserted, and the noise of the operational amplifier OA and the whole passive network including the diode bridge D_1 to D_4 is measured. The blocking capacitor C_{c3} prevents the signal from penetrating the control voltage circuits, U_c .

The change of the operating point of the bridge is caused by the change of the current I_{ctrl} . This current is determined by the magnitude of the control voltage U_c and the magnitude of the resistors R_{bias} . Since the same VA characteristics of diodes D_1 to D_4 are assumed here as well, they are replaced with identical dynamic resistors r_d (it is not possible to use

a substitute resistor of the whole bridge r_{dfinal} , because the integrity of the bridge is disrupted by the resistors R_{bias}). Based on such a linearized model, the following applies:

$$A = \dots \text{(see in Appendix IX)} \quad (25)$$

Relationship (25) can, of course, be further simplified. On the other hand, freestanding fractions with the variable r_d are advantageous and allow easy analysis for $r_d \rightarrow \infty \Omega$, or $r_d \rightarrow 0 \Omega$:

$$\lim_{r_d \rightarrow \infty} A = -\frac{1}{R_{amp1} \left(\frac{1}{R_{amp2}} \right)} = -\frac{R_{amp2}}{R_{amp1}} \quad [/] \quad (26)$$

$$\begin{aligned} \lim_{r_d \rightarrow 0} A &= -\frac{1}{R_{amp1} \left(\frac{1}{R_{amp2}} + \frac{1}{R_{d2}} \right)} = \\ &= -\frac{R_{amp2}R_{d2}}{R_{amp2}+R_{d2}} = -\frac{R_{amp2}||R_{d2}}{R_{amp1}} \quad [/] \quad (27) \end{aligned}$$

Since relationship (26) coincides with relationship (17) and relationship (27) coincides with relationship (18), it can be stated that due to the resistors R_{bias} , there are no shifts in the extreme limits of voltage transfer A . However, the relationship (25) for voltage transfer A differs from the original relation (16), thanks to the R_{bias} resistors. As will be shown below, the relationships for NG noise gain or transmission from the feedback network β , also differ. However, it still holds that if the diode bridge is disconnected ($I_{ctrl} = 0 \text{ A}$), the high dynamic resistance of the diodes r_d prevents any penetration of noise into the signal path, whether it is the noise voltage of the resistors R_{d1} , R_{d2} , R_{bias} , or the noise of the control voltage U_c . The shot noise of the diodes, $2q(I_{ctrl}/2)$, is also zero because the control current I_{ctrl} does not flow through. If $R_{bias} \rightarrow \infty \Omega$, is considered, relation (25) is simplified to the form of relationship (16).

The stage design is based on the limit transmissions defined by expressions (26) and (27). The basis is the required maximum gain A given by relationship (26) or relationship (1). For the values of resistors R_{amp1} and R_{amp2} , the required minimum gain and acceptable value R_{d2} must be taken into account. If the smallest gain is to be equal to 1, it is necessary to take into account the equality $R_{amp1} = R_{d2}$. It will now depend on the acceptable value of the stage load ($R_{d1} + R_{d2}$ to ground) and the allowable minimum input resistance of the stage (R_{amp1} to virtual ground).

Another important step is the design of the resistance divider R_{d1} and R_{d2} . Its division ratio is based on the required maximum output amplitude of the stage and the allowable distortion. It is advantageous to select this division ratio to be as large as possible, however, it is limited by relationship (12). The value of the resistors, R_{bias} , is based on the maximum value of the current I and the range of the control voltage U_c . The coupling capacities C_{c1} and C_{c2} , together with the output resistance of the divider R_{d1} and R_{d2} and the dynamic resistance R_d , determine the lower cut-off frequency f_L , from which the gain controller begins to be used. As the dynamic resistance R_d decreases, the lower cut-off frequency f_L shifts

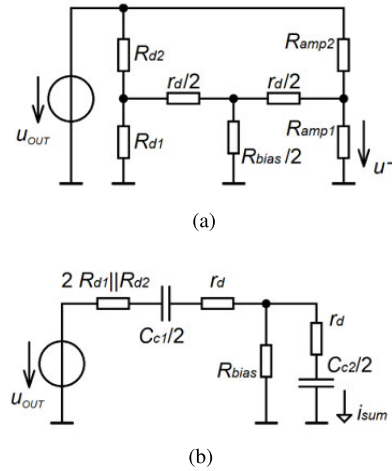


FIGURE 6. Linearized models for calculation of a) upper limit frequency f_{CL} , b) lower limit frequency f_L .

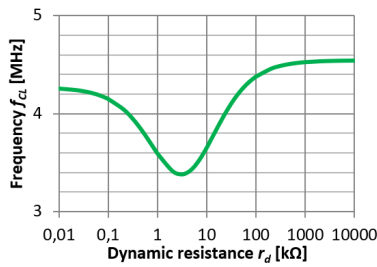
upwards. It is greatest at zero value of dynamic resistance. The sizes of the coupling capacities must therefore be chosen so as to ensure a maximally flat frequency response characteristic in the range of the transmitted band.

The task of the described amplifier was to connect to a photoelectric converter and provide the necessary amplification for processing a Manchester encoding data stream. The bandwidth of the amplifier was set at 20 kHz and 4.5 MHz. The LT1360 (Analog Devices, Norwood, MA, USA, 1994) with the upper cut-off frequency $f_T = 50 \text{ MHz}$ was chosen as the operational amplifier OA. At the required maximum gain $A = 10$ of one stage, the transmission of the feedback network according to relationship (2) was equal to 1/11 and the upper cut-off frequency f_{CL} of the closed loop according to relationship (3) was equal to 4.5 MHz. Other component values were designed as follows: $R_{amp1} = 2\text{k}\Omega$, $R_{amp2} = 22\text{k}\Omega$, $R_{d1} = 100\Omega$, $R_{d2} = 1\text{k}\Omega$, $R_{bias} = 2\text{k}\Omega$, $C_{c1} = 100\text{nF}$, $C_{c2} = 100\text{nF}$, $C_{c3} = 100\text{nF}$, D₁ to D₄... 1N4148 (Vishay, Malvern, PA, USA, 2017) (measured and thermally bonded). As can be seen, the optimal value of the resistor R_{amp1} was chosen to be 2k Ω . With the capacitance $C_{sum} = 5 \text{ pF}$, the pole frequency is then $f_{pool} = 14.5 \text{ MHz}$, relation (13), which is already beyond the upper cut-off frequency of the closed loop f_{CL} and will not affect its transmission (the closed loop transmission function will be without resonant cant). Because the impedance in the bridge control path was small ($2 \times R_{bias} = 2\text{k}\Omega$), the upper cut-off frequency f_{CL} of the closed loop with the gain change was not constant, but fluctuated.

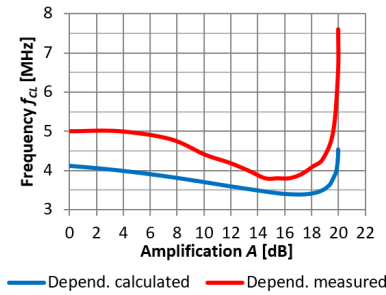
V. PARAMETERS OF THE BASIC VGA CIRCUIT DESIGN

A. FREQUENCY CHARACTERISTICS

First, the frequency response characteristic of the basic VGA circuit design according to Fig. 5 will be investigated. The linearized model from Fig. 6(a) will be used to investigate the dependence of the passive network β transmission on the dynamic resistance of the diodes r_d .



(a)



(b)

FIGURE 7. Graph of the dependence of the upper cut-off frequency of the closed loop f_{CL} on a) dynamic resistance of diodes r_d , b) on amplification A .

The transmission of the feedback network β is given by the following expression:

$$\beta = \frac{u^-}{u_{out}} = \dots (\text{see in Appendix IX}) \quad (28)$$

The transmission of the feedback network β was found using the nodal voltage method, while making use of the simplification resulting from the symmetry of the bridge. The graph of the transmission dependence of the feedback network β on the dynamic resistance of the diodes r_d is shown on 7(a), and the graph of the dependence of the upper cut-off frequency f_{CL} on the gain A is shown in Fig. 7(b).

The actual closed loop response of the operational amplifier and the feedback network shows a slight resonant hump, see Fig. 8(a), therefore the measured values of the upper cutoff frequency of the closed-loop f_{CL} depending on the gain A are slightly higher than calculated.

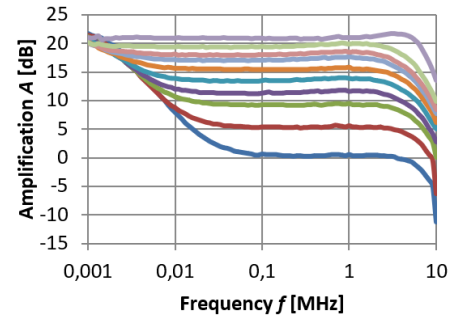
As already shown above, the R_{bias} resistors affect the transmission of the feedback network β and make it frequency-dependent again. The aim is therefore to eliminate R_{bias} resistors by increasing their resistance ($R_{bias} \rightarrow \infty \Omega$). Relationship (28) is rewritten as follows:

$$\lim_{R_{bias} \rightarrow \infty} \beta = \dots (\text{see in Appendix IX}) \quad (29)$$

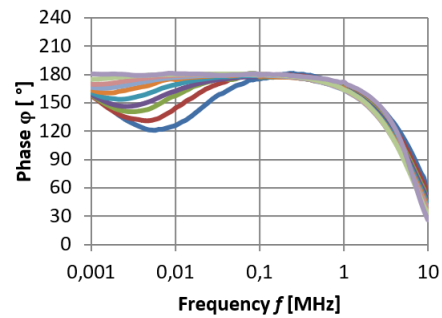
If, moreover, condition (23) is met, relationship (29) is further simplified:

$$\lim_{R_{bias} \rightarrow \infty} \beta = \dots (\text{see in Appendix IX}) \quad (30)$$

Relationship (30) coincides with the inverse value of relationship (24). Thus, if the values of the resistors



(a)



(b)

FIGURE 8. Measured a) amplitude and b) phase frequency response characteristic of a VGA with a diode bridge.

are $R_{bias} \rightarrow \infty \Omega$ and if condition (23) is satisfied, the transmission β and with it the upper cutoff frequency of the closed loop f_{CL} remain constant despite the change in the gain A (or also the dynamic resistance r_d).

It remains to examine the lower cut-off frequency f_L . For this, we will use the linearized model, according to Fig. 6(b). The ratio of the input voltage u_{OUT} and the output current i_{sum} is given by the following expression:

$$\frac{u_{OUT}}{i_{sum}} = \dots (\text{see in Appendix IX}) \quad (31)$$

The impedance of u_{OUT}/i_{sum} was found using the loop currents method, while taking advantage of the simplification resulting from the symmetry of the bridge. A further procedure consists in dividing the expression into real and imaginary parts, whereas we are looking for their equality for the given r_d . This occurs at the lower cut-off frequency f_L . The graph of the dependence of the lower cut-off frequency f_L on the dynamic resistance of the diodes r_d is shown in Fig. 9. The lower cut-off frequency values found only apply to the gain controller. The lower cut-off frequency of the whole amplifier is then slightly higher, because the resulting transmission A is determined by a parallel combination of the

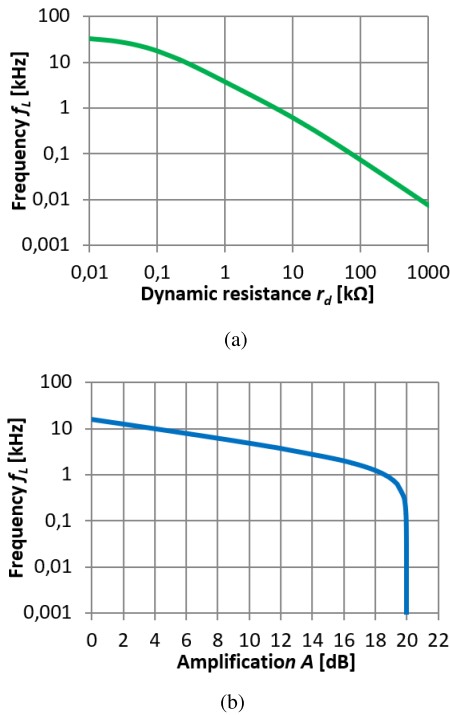


FIGURE 9. Graph of the dependence of the lower cut-off frequency f_L on the a) dynamic resistance of the diodes r_d and b) on the amplification A .

gain controller and the basic feedback network R_{amp2}/R_{amp1} of the operational amplifier.

B. DISTORTION

Furthermore, it is necessary to investigate the distortion introduced by the diode bridge in the case when $I_{ctrl} \neq 0$ A. The diode attenuator according to Fig. 10(a) serves as the basis of the theoretical analysis. If the voltage on diode D_1 drops, its current, i_1 , also drops, but not by as much as would correspond to a drop in the movement of the operating point along the tangent, 10(b). Thus, in the case of a single diode, a larger current, i_{div} , will flow through the resistor R than would flow with a resistive load. However, there is a second diode here. The voltage on diode D_2 increases. With it, the current i_2 also increases and again by more than would correspond to the increase in the movement of the working point along the tangent. Since more current i_2 flows out of the node through the diode D_2 than would be the case with a resistive load, the excess current i_1 is compensated, which in turn flows into the node through diode D_1 . The change in the current, i_{div} , will therefore be very close to the movement of the operating point tangentially, since the currents through the diodes i_1 and i_2 are partially compensated. It applies that:

$$i_{div} = i_2 - i_1 \tag{32}$$

The current through the diode is expressed by relation (4) and, as the measurements prove, the exponential dependence of the current I on the voltage U is very close to reality. To obtain the frequency spectrum, it will therefore be

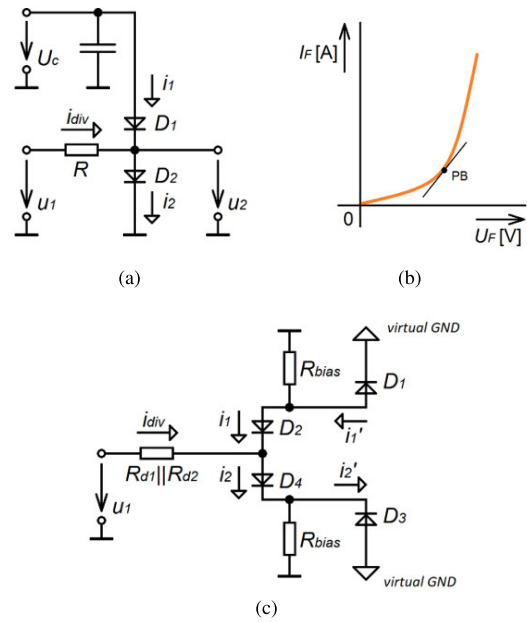


FIGURE 10. Wiring of the diode divider, b) VA diode characteristics, c) substitute model of bridge for alternating signals.

sufficient to express the exponential function e^x using a power series, Taylor expansion, around the point x_0 :

$$f(x) = f(x_0) + \frac{f'(x_0)}{1!} (x-x_0) + \frac{f''(x_0)}{2!} (x-x_0)^2 + \dots \tag{33}$$

The condition is the existence of derivatives of the function $f(x)$ of all orders. For simplicity, we further consider the development of the exponential function e^x around the point $x_0 = 0$. This leads to the notation:

$$e^x = 1 + \frac{x}{1!} + \frac{x^2}{2!} + \frac{x^3}{3!} + \dots \tag{34}$$

If the value is $x > 0$, notation (34) remains unchanged. If the value is $x < 0$, the relationship (34) is rewritten as follows:

$$e^{-x} = 1 - \frac{x}{1!} + \frac{x^2}{2!} - \frac{x^3}{3!} + \dots \tag{35}$$

To expression (4), then $x = U/2U_T$. The voltage change u_2 of the diode divider according to Fig. 10(a) represents a positive voltage change $U(x > 0)$, for one diode and a negative voltage change $U(x < 0)$ for the other diode. Based on expression (32), the current through resistor R is found as the difference between the currents of both diodes:

$$\begin{aligned} I_{div} - I_2 - I_1 &= I_0 (e^x - e^{-x}) \\ &= 2I_0 \sinh x = 2I_0 \sinh \frac{U}{2U_T} \end{aligned} \tag{36}$$

Relationship (36) can also be rewritten using the power series (34) and (35):

$$\begin{aligned} I_{div} - I_2 - I_1 &= I_0 (e^x - e^{-x}) \\ &= 2I_0 \left(\frac{x}{1!} + \frac{x^3}{3!} + \frac{x^5}{5!} + \dots \right) \end{aligned} \tag{37}$$

It follows from relationship (37) that, due to the joint action of both diodes, even members of the development and hence even harmonic components of the current have been compensated, which is desirable. The basic opinion regarding the distortion introduced by, for example, the third harmonic component is then obtained as the ratio of the cubic and the first term of the power series:

$$K_3 \approx \frac{\frac{x^3}{3!}}{\frac{x}{1!}} = \frac{x^2}{3!} = \frac{1}{3!} \left(\frac{U}{2U_T} \right)^2 \quad (38)$$

The voltage amplitude u_2 is substituted at the output of the divider at the voltage U . Although the derivation given above is only approximate, it achieves a fair agreement with reality.

As will be shown below, the diode bridge distortion relationships can be obtained with only a small modification of the diode divider distortion relationships. In both cases, the distortion of the same current, i_{div} , which is transformed into an output, is investigated. In the case of a diode divider, it is a transformation to the output voltage u_2 , i.e., $u_2 = u_1 - i_{div}R$, and in the case of a diode bridge, we are dealing with a transformation to the output voltage of the operational amplifier u_{OUT} , i.e., $u_{OUT} = (i'_1 + i'_2)R_{amp2} \approx i_{div}R_{amp2}$. In the case of a diode divider, Fig. 10(a), there is only one diode between its center and the signal ground. In the case of a diode bridge, Fig. 10(c), there is a series array of two diodes between its center and the signal ground (the resistance R_{bias} does not apply due to higher ohmic values). If the diodes were polarized identically (cathode-anode and cathode-anode), this would be a standard series arrangement of diodes, which in relation (4) would result in a doubling of the denominator of the exponent. According to relationship (38), the allowable signal amplitude for a given distortion would increase $2\times$, or potentially, for the given signal amplitude, the distortion would decrease $4\times$. However, because the considered diode bridge polarizes the diodes inversely (cathode-anode and anode cathode), the characteristics of the diodes are inverse (concave and convex), the nonlinearities are compensated and the resulting distortion is even smaller. Therefore, much better results can be expected than with a diode divider, and relationship (38) can be adjusted as follows:

$$K_3 < \frac{1}{3!} \left(\frac{\frac{U}{2}}{2U_T} \right)^2 = \frac{1}{3!} \left(\frac{U}{4U_T} \right)^2 \quad (39)$$

Relationship (39) can be verified using the measured data. Let the voltage amplitude at the output of the divider R_{d1} and R_{d2} be 141 mV (this corresponds to $u_{OUT} = 1.41$ V at the output of the operational amplifier!) and let the dynamic resistance of the diodes r_d be much greater than the output resistance of the divider R_{d1} and R_{d2} (90Ω). It then holds that $U = 141$ mV, and the distance I. and III. of the harmonic component, according to relationship (39), is equal to 10 dB. Under the same conditions (corresponding to a gain of 10 dB), a distance between I. and III. of the harmonic component of around 24 dB is measured. It can be seen that the measured

value is 14 dB better than the theoretical one and the inequality in relation (39) is satisfied.

Furthermore, if we let the voltage amplitude at the output of the divider R_{d1} and R_{d2} again be 141 mV, but the dynamic resistance of the diodes r_d is the same as the output resistance of the divider R_{d1} and R_{d2} (90Ω), it then holds that $U = 70$ mV, and the distance between I. and III. of the harmonic component, according to equation (39) is equal to 22.5 dB. Under the same conditions (corresponding to a gain of 0 dB), a distance of around 32.5 dB between I. and III. of the harmonic component is measured. It can be seen that the measured value is 10 dB better than the theoretical one and the inequality in relation (39) is again met.

As follows from the theoretical analysis or from the measured data, the diode bridge can process a relatively large signal with little distortion. In general, a minimum better distance of 10 dB between I. and III. harmonic components can be expected than the value calculated according to relationship (39). The higher harmonic components found are only odd, while the most important ones are III. and V.

Furthermore, measurements showed that the size of the III. and the V. harmonic component decreases rapidly when decreasing the I. harmonic component of the output voltage u_{OUT} . If at the gain $A = 10$ dB and the magnitude of the I. harmonic component was 0 dBV (effective value 1 V) and the distance of the III. and V. harmonic components -24 and -42 dB, then a decrease of the I. harmonic component by half to -6 dBV (effective value 0.5 V) increased the distance between the III. and V. harmonic components to -38 and -66 dB, Fig. 12. If the reduction of the output voltage u_{OUT} was undesirable, the distortion could be reduced only by adjusting the division ratio R_{d1}, R_{d2} .

C. NOISE ANALYSIS

An alternative scheme according to Fig. 13 was used to calculate the noise voltage at the output of the VGA amplifier. The contributions of the individual noise sources were recalculated to the output (RTO... Referred to Output), so that they could be easily verified in the end by measuring the output noise voltage u_n . The noise investigation was performed in two steps.

First, the OA operational amplifier itself with the R_{amp1} a R_{amp2} feedback network was examined. For this purpose, the jumper S was removed. In the above configuration, the gain A was at maximum with a value of 20 dB. An overview of noise sources, their sizes and transformations to output is given in Table 1. When calculating the thermal noise of resistors R_{amp1} a R_{amp2} , a room temperature of $T = 300$ K was considered. Further, an infinitely large gain of the operational amplifier OA, $K_{21} \rightarrow \infty$, hence $K_{CL} \approx 1/\beta = NG$, was considered. The total voltage power spectral noise density at the output of the operational amplifier was found as the sum of all values of the last column in Table 1, and was $14.153 \cdot 10^{-15} \text{ V}^2 \cdot \text{Hz}^{-1}$. To be able to compare this value with the measurement, it was multiplied by the bandwidth of the filter used by 300 Hz, its square root taken, and converted to a decibel measure. The

TABLE 1. Overview of noise sources of the operational amplifier OA and its feedback network R_{amp1} , R_{amp2} .

Noise source	Size	Mathematical transformation	Calculation of voltage power spectral output density	Voltage power value of spectral output density
$\overline{e_n}$ (LT1360)	$9 \cdot 10^{-9}$ [V · Hz ^{-1/2}]	$K_{CL}^2(f) = \left[\frac{1}{\beta(f) + \frac{1}{K_{21}(f)}} \right]^2$	$\overline{e_n}^2 K_{CL}^2(f)$	$9.8 \cdot 10^{-15}$
$\overline{J_a}$ (LT1360)	$0.9 \cdot 10^{-12}$ [A · Hz ^{-1/2}]	$K_I^2(f) = R_{amp2}^2$	$\overline{J_a}^2 K_I^2(f)$	$392 \cdot 10^{-18}$
$\overline{e_{nRamp1}}$ (2k2)	$6 \cdot 10^{-9}$ [V · Hz ^{-1/2}]	$K_{Ramp1}^2(f) = \left(\frac{R_{amp2}}{R_{amp1}} \right)^2$	$\overline{e_{nRamp1}}^2 K_{Ramp1}^2(f)$	$3.6 \cdot 10^{-15}$
$\overline{e_{nRamp2}}$ (22k)	$19 \cdot 10^{-9}$ [V · Hz ^{-1/2}]	$K_{Ramp2}^2(f) = 1$	$\overline{e_{nRamp2}}^2 K_{Ramp2}^2(f)$	$361 \cdot 10^{-18}$

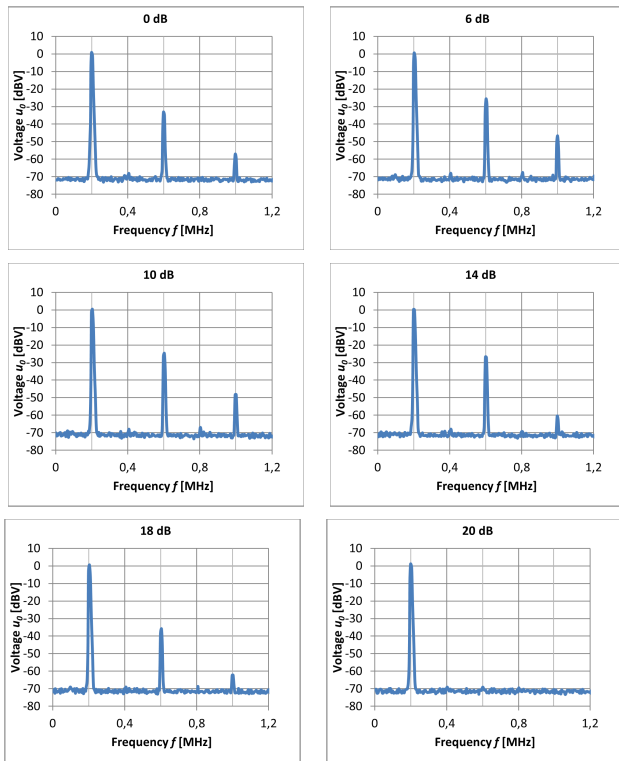


FIGURE 11. Measured higher harmonic components at the output of the operational amplifier and gain 0 to 20 dB (Rohde & Schwarz generator 826.4517.52, Tektronix 7L5 spectrum analyzer).

value adjusted in this way was -113.7 dBV. A Tektronix 7L5 spectrum analyzer was used for practical verification. At the mentioned filter bandwidth of 300 Hz, the measured values were around -115 dBV, which is a good agreement with the calculation.

In the second step, the entire VGA amplifier was investigated for various amplifications A . A jumper S was connected. The dominant source of noise remained the voltage noise of the operational amplifier $\overline{e_n}$, which was transformed to the output by the factor K_{CL} according to the current value of the transmission feedback network β , Table 2. The contribution of the thermal noise $\overline{e_{nRamp1}}$ of the resistor R_{amp1} was transformed to the output by the factor K_{Ramp1} according to the current value of the gain A , Table 2. The contribution

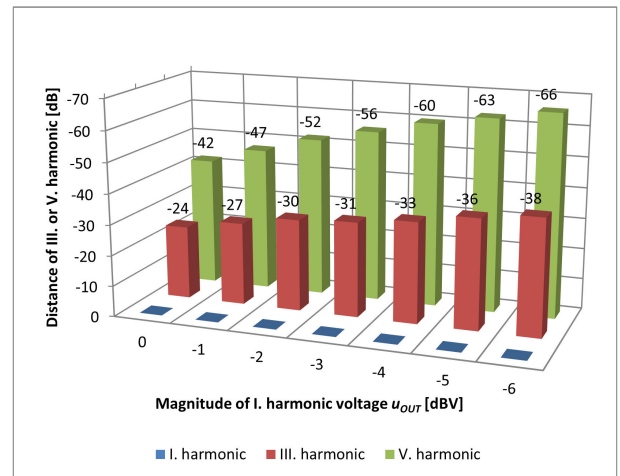


FIGURE 12. Measured distances III. and V. of harmonic components depending on the size of the I. harmonic component at gain $A = 10$ dB.

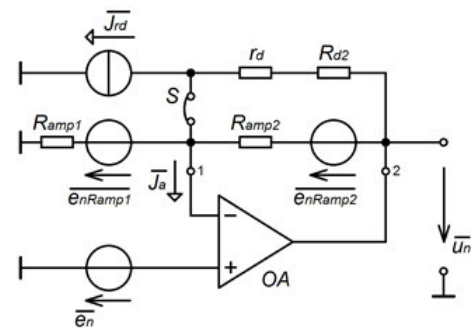


FIGURE 13. Substitute diagram of VGA amplifier showing noise sources.

of the thermal noise $\overline{e_{nRamp2}}$ of the resistor R_{amp2} could be overlooked, as well as the contribution of the current noise of the operational amplifier $\overline{J_a}$, see the values in Table 1. Finally, the current spectral density of the diode bridge $\overline{J_{rd}}$ and its transformation to the output were investigated by means of the factor $K_I = R'_{amp2}$. For each gain A , the magnitude of the replacement resistance $R'_{amp2} = A \cdot R_{amp1}$ and the magnitude of the current through diode I were determined, from which the current spectral density (shot noise) of the whole diode bridge was calculated:

$$\overline{J_{rd}^2} = 4 \cdot 2qI = 8qI [A^2 Hz^{-1}] \quad (40)$$

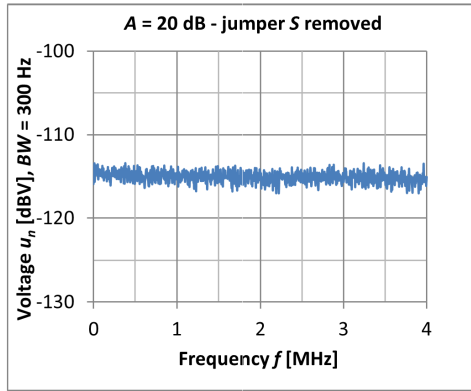


FIGURE 14. Measured spectral density of the noise voltage u_n of the amplifier according to Fig. 13.

The total voltage power spectral noise density at the output of the operational amplifier ($\overline{u_n^2}$) is equal to the sum of the values of the highlighted rows of the table according to the respective gain A . A Tektronix 7L5 spectrum analyzer was used for verification. The measured spectral noise densities are shown in Fig. 15, from which a good agreement with the calculations can be seen. The actual noise level is slightly lower because the filtering effects of the parasitic capacitances of the connections were not taken into account in the calculations.

By comparing Fig. 14 and Fig. 15 for gain $A = 20$ dB, it can be concluded that the diode bridge at zero control voltage U_c does not contribute to the resulting noise. If the control voltage U_c varies, from zero, the shot noise of the bridge diodes increases, however, this increase is again compensated by the decrease in the thermal noise contribution of the resistor R_{amp1} . The total voltage power spectral density of the noise at the output of the operational amplifier more or less copies the contribution of its voltage noise depending on the noise gain NG , resp. on the inverse value of the feedback network transmission β .

VI. WIRING OF A VGA WITH A CONSTANT BANDWIDTH

As can be seen from relationship (30), if the values of the resistors are $R_{bias} \rightarrow \infty \Omega$ and if condition (23) is met, the transmission β , and with it the upper cut-off frequency f_{CL} , remain constant despite the change in gain A . Fulfilling condition (23) is relatively simple, it is only necessary to maintain the same ratios of R_{amp2}/R_{amp1} and R_{d2}/R_{d1} . However, meeting the $R_{bias} \rightarrow \infty \Omega$ condition is more challenging. High impedance in the control path of the diode bridge can only be ensured by means of a current source. The first option of providing a current source is to use a high voltage source in series with a high-ohmic resistor. Because the implementation of a high voltage source is not a trivial matter, simpler solutions are sought. In microelectronic circuits, it is more advantageous to use high impedances of the collector circuit of a transistor, e.g., Wilson current mirrors, see Fig. 16(a), while in discrete circuits, on the other hand, bootstrap type

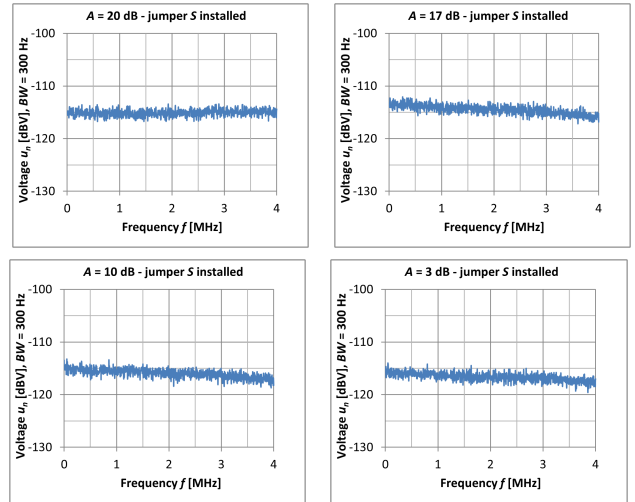
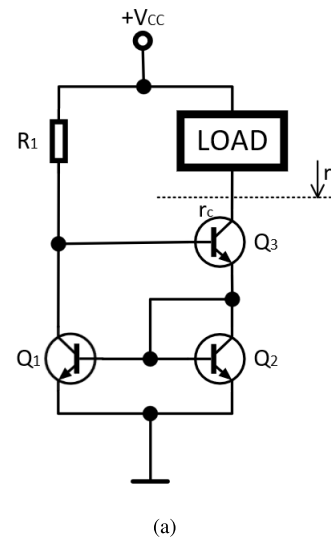
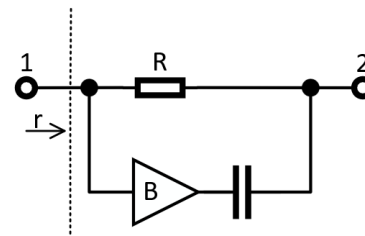


FIGURE 15. The measured spectral density of the noise voltage u_n of the amplifier according to Fig. 13.



(a)



(b)

FIGURE 16. Current source executed using a) Wilson current mirror, b) bootstrap – type feedback structure.

feedback structures are more advantageous, see Fig. 16(b). In both cases, current sources with a large internal resistance “ r ” can be achieved.

In the described VGA amplifier, the current sources were implemented using feedback structures of the “bootstrap”

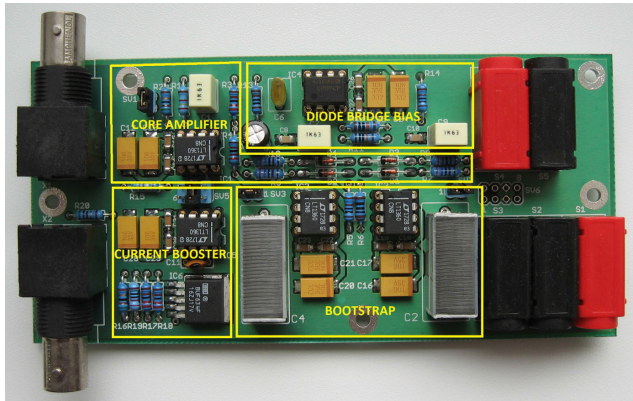


FIGURE 18. VGA amplifier with constant bandwidth (current-controlled diode bridge).

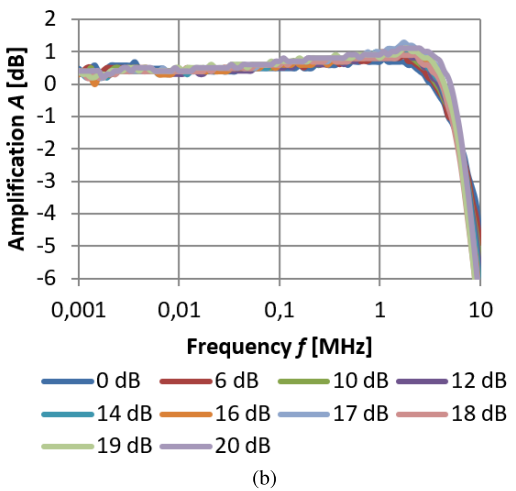
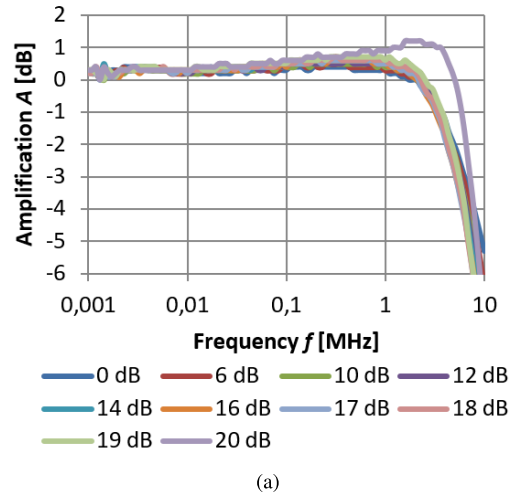


FIGURE 20. Normalized amplitude frequency response characteristic of a VGA a) without current sources, b) with current sources on the control path of the diode bridge.

VII. VGA GAIN CONTROL

Next, the dependence of the gain A on the dynamic resistance of the diodes r_d and on the control current I_{ctrl} will be investigated. As shown in the previous chapter, the constant bandwidth of a VGA amplifier can be ensured by the limit $R_{bias} \rightarrow \infty \Omega$, at which the relationship (25) for amplification, A , changes into the form of relationship (16). Although in relationship (25) we work with the dynamic resistance r_d of one diode and in relation (16) with the dynamic resistance r_{dfinal} of the whole bridge, they are identical in value; they will therefore not be further distinguished and will be replaced by the dynamic resistance r_d of one diode. Let the dynamic resistance r_d of one diode be expressed as an independent variable x , let K , L , M and N be constants, and finally let the gain be a dependent variable $A(x)$. Then relationship (16) can be rewritten in the following form:

$$A(x) = N \cdot \frac{Kx + L}{Kx + M} [/] \tag{41}$$

It is clear from expression (41) that this is a rational polynomial function in which the relationship between the

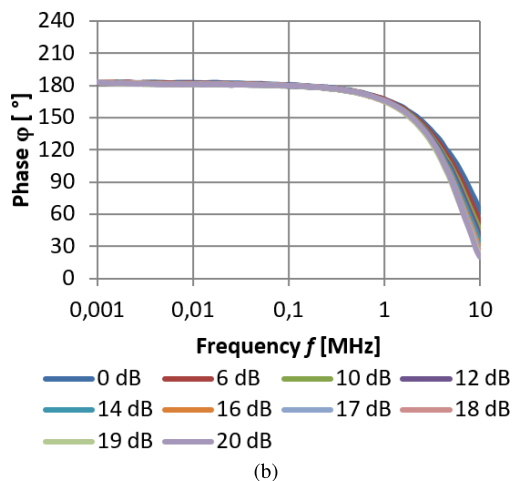
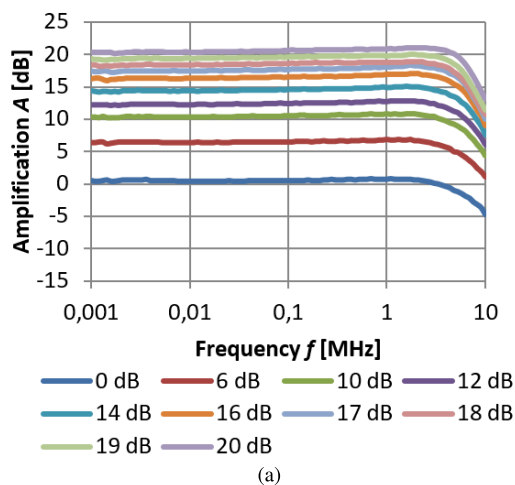
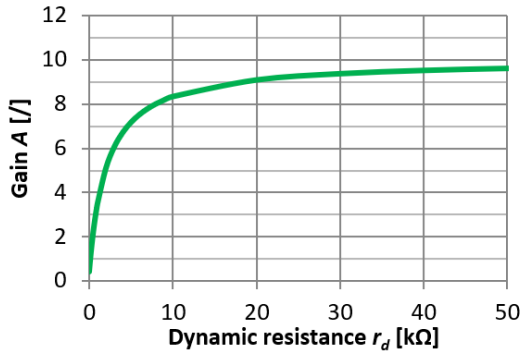
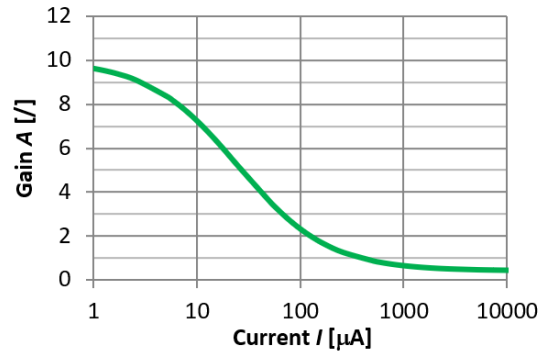


FIGURE 19. Measured a) amplitude and b) phase frequency characteristic of VGA with current sources.

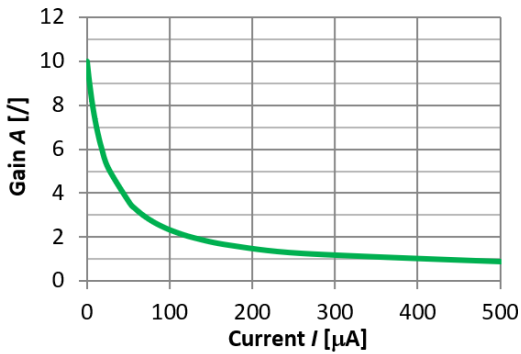
Fig. 20(a), with current sources in Fig. 20(b). A comparison of the figures shows that the current sources provide a constant gain-independent bandwidth, the magnitude of which corresponds to the basic wiring of the inverting operational amplifier according to relationships (1), (2) and (3).



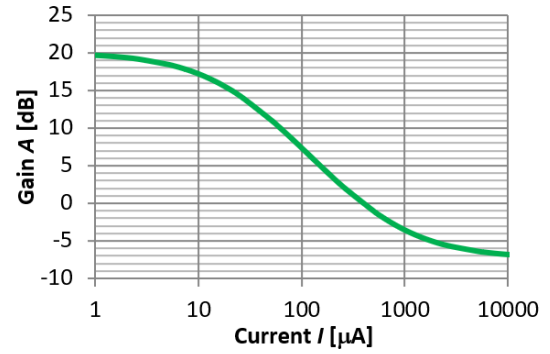
(a)



(a)



(b)



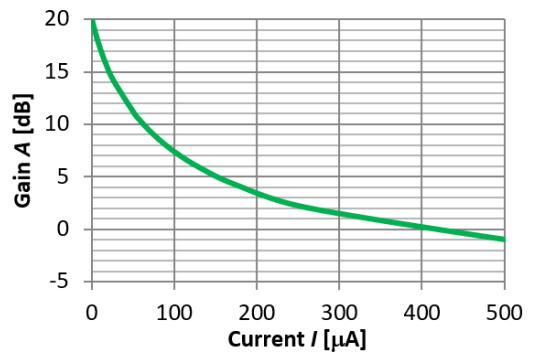
(b)

FIGURE 21. Gain, A , dependence on: a) dynamic resistance diodes r_d , or b) on diode current I .

independent and dependent variables is generally nonlinear. This fact is confirmed by the graph in Fig. 21(a). The situation does not improve even if the dynamic resistance r_d is expressed by the current through the diode I according to relationship (7), or by the control current $I_{ctrl} = 2I$, see Fig. 21(b).

A logarithm of the current through the diode I or the control current I_{ctrl} could be a certain help, see Fig. 22(a). However, even with this adjustment, only the interval between the gain value 3 to 7 will be linear. If a linear dependence of the gain A on the control current I_{ctrl} in the whole range is required (e.g., for modulators), a function converter would have to be used. On the other hand, VGA with a diode bridge works well in automatic gain control loops, where the output signal level is compared to an external normal and the amount of gain is provided by feedback (but the stability of the loop over the entire range of possible gains must be checked). For completeness, the dependence of the gain A expressed in dB on a logarithmic change of the current I , Fig. 22(b), and a linear change of the current I , Fig. 22(c), is also added.

The presented VGA amplifier wiring of Fig. 17 is designed for AC signal processing. It uses several RC cells. Therefore, the following is a description of these cells, including possible circuit alternatives.



(c)

FIGURE 22. Dependence of gain, A , where a) is the gain plotted as a ratio on a logarithmic change of current through the diode I , where b) is the gain plotted in the decibel level on a logarithmic change of current through the diode I , and where c) is the gain plotted in the decibel level on a linear change of current through the diode I .

VIII. VGA OFFSET ELIMINATION

One of the critical parameters of the amplifier is its bandwidth. The bandwidth limitation from above is usually determined by the technology used. The bottom bandwidth limitation, in turn, is due to the AC coupling between the stages, which is based on the need to separate the DC voltages and currents of the operating points of the active components and the AC voltages and currents of the signals being processed. In some cases, AC couplings are irreplaceable and a decrease in the lower cut-off frequency is only possible with an increase in the values of resistors, capacitors, or a combination of both. If DC signal processing is also required,

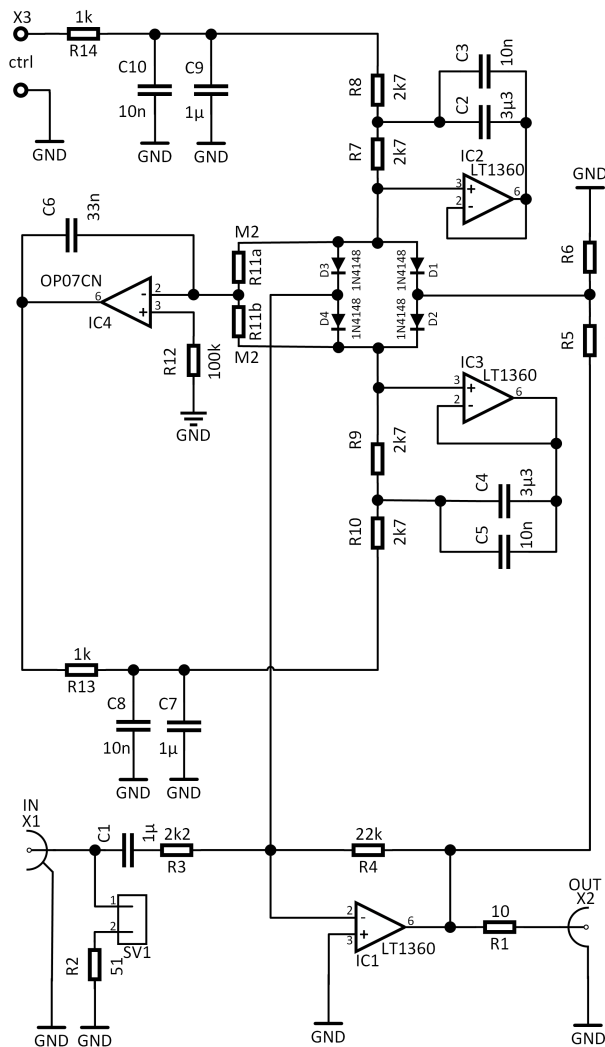


FIGURE 23. Alternative wiring of a constant bandwidth VGA amplifier without capacitors C11 and C12.

or if it is an integrated solution excluding large time constants of RC cells, parasitic DC components must be eliminated by appropriate design.

The first cell, $R3 \cdot C1$, forms the upper pass. Capacitor C1 separates any DC component of the previous stage. If the input signal does not contain a DC component, capacitor C1 can be omitted. Decreasing the value of capacitor C1 is possible only with increasing the resistances of the passive network of the operational amplifier IC1 and the dynamic resistance of the bridge diodes. The other high-pass filters are cells $(R8 \parallel R7) \cdot (C2 \parallel C3)$ or $(R10 \parallel R9) \cdot (C4 \parallel C5)$ providing current excitation of the diode bridge. In the case of the discrete variant according to Fig. 17, capacitors C2||C3 and C4||C5 cannot be omitted, it is only possible to reduce their value, at the cost of increasing the lower cut-off frequency, from which the “bootstrap” begins to apply, i.e., an increase in resistance in the bridge control path. If it were a microelectronic application of the described VGA amplifier, it would use, for example, Wilson current mirrors, which do not need capacitors for their function.

The $R8 \cdot (C9 \parallel C10)$ and $R10 \cdot (C7 \parallel C8)$ cells form a low pass filter and prevent the signal from penetrating the control path of the diode bridge. The values of the larger capacitors C7 and C9 are not critical and can be radically reduced or omitted if the impedance of the output IC4 or the external control source is low. If C7 to C10 are fitted, it is necessary to separate the output IC4 with R13, or an external source with R14.

The final cell $(C11 \parallel C12) \cdot r_{dfinal}$ again forms the upper pass. Capacitors C11 and C12 are necessary in the current wiring according to Fig. 17 in order to separate the direct currents of the operational amplifiers IC1 and IC4. If they were not fitted, the zero DC component at the output of the diode bridge would be provided not by IC4 but by IC1, and the supply voltage of the bridge would not be symmetrical. There is no voltage on the capacitors during operation, so the smallest possible types can be used. The correct function of IC1 and IC4 circuits can be ensured even without the use of capacitors C11 and C12. A functional alternative may be to divide resistor R11 into a pair of resistors R11a and R11b, as shown in Fig. 23. The integration capacitor C6 will then be charged until the supply voltage to the bridge is symmetrical.

IX. CONCLUSION

The scope of the paper was the design and description of the properties of an amplifier with gain-control, which was designed with regard to the following objectives: 1) a simple structure suitable not only for integrated solutions but also for production with discrete components, 2) a design based on an operational amplifier with voltage feedback, enabling continuous change of gain using one element (suitable for AGC systems), 3) constant amplifier bandwidth despite gain change, 4) a separate design and independent selection of amplifier and gain controller parameters. The solution found came from the inverting circuit of an operational amplifier, whose passive resistor network was extended by another feedback element T, partly formed by a diode bridge. Based on the model for alternating signals, relations were derived for the amplification of structure A, and for the transmission of the feedback network β . Furthermore, conditions were derived ensuring a constant value of β and thereby even the bandwidth of the amplifier through the change of gain. The first presented circuit solution was assembled to verify the operation of the VGA amplifier and measure its parameters. The control circuit of the diode bridge proved to be a problem, which, due to its low impedance, made it impossible to achieve a constant bandwidth. The second presented solution increased the impedance in the bridge control circuit by means of current sources and thus fulfilled the condition leading to a constant bandwidth. The current sources were accomplished with a bootstrap feedback structure, and the use of a Wilkinson current mirror was proposed for the microelectronic variant. Both presented solutions were documented by measuring amplitude and phase frequency characteristics. A detailed analysis of the distortion and noise ratios in the circuit was also presented. It was shown that if the maximum

$$\begin{aligned}
 NG &= \frac{1}{\beta} = \frac{r_{dfinal}(R_{d1} + R_{d2})(R_{amp1} + R_{amp2}) + R_{d1}R_{d2}(R_{amp1} + R_{amp2}) + R_{amp1}R_{amp2}(R_{d1} + R_{d2})}{R_{amp1}(r_{dfinal}R_{d1} + r_{dfinal}R_{d2} + R_{d1}R_{d2} + R_{d1}R_{amp2})} [1] \\
 NG &= \frac{r_{dfinal}(R_{d1} + R_{d2})(1 + \frac{R_{amp1}}{R_{amp2}}) + R_{d1}R_{d2}(1 + \frac{R_{amp1}}{R_{amp2}})R_{amp1}(R_{d1} + R_{d2})}{R_{amp1}(\frac{r_{dfinal}R_{d1}}{R_{amp2}} + \frac{r_{dfinal}R_{d2}}{R_{amp2}} + \frac{R_{d1}R_{d2}}{R_{amp2}} + R_{d1})} [1] \\
 NG &= \frac{r_{dfinal}(R_{d1} + R_{d2})\frac{R_{amp1}+R_{amp2}}{R_{amp1}} + R_{d1}R_{d2}\frac{R_{amp1}+R_{amp2}}{R_{amp1}} + R_{amp2}(R_{d1} + R_{d2})}{r_{dfinal}R_{d1} + r_{dfinal}R_{d2} + R_{d1}R_{d2} + R_{d1}R_{amp2}} \\
 &= \frac{r_{dfinal}\frac{R_{d1}+R_{d2}}{R_{d1}}\frac{R_{amp1}+R_{amp2}}{R_{amp1}} + r_{d2}\frac{R_{amp1}+R_{amp2}}{R_{amp1}} + r_{amp2}\frac{R_{d1}+R_{d2}}{R_{d1}}}{r_{dfinal}\frac{R_{d1}+R_{d2}}{R_{d1}} + R_{d2} + R_{amp2}} \\
 \beta &= \frac{u^-}{u_{out}} \\
 &= \frac{\frac{R_{d2}}{2} \left(\frac{1}{R_{d1}} + \frac{1}{R_{d2}} + \frac{2}{r_d} \right) \left(\frac{2}{r_d} + \frac{1}{R_{bias}} \right) + \frac{R_{amp2}}{r_d^2} - \frac{R_{d2}}{r_d^2}}{\frac{R_{d2}}{2} \left(\frac{1}{R_{d1}} + \frac{1}{R_{d2}} + \frac{2}{r_d} \right) + \left(\frac{2}{r_d} + \frac{1}{R_{bias}} \right) R_{amp2} \left(\frac{1}{R_{amp1}} + \frac{1}{R_{amp2}} + \frac{2}{r_d} \right) - \frac{R_{amp2}R_{d2}}{r_d^2} \left(\frac{1}{R_{d1}} + \frac{1}{R_{d2}} + \frac{2}{r_d} \right) - \frac{R_{amp2}R_{d2}}{r_d^2} \left(\frac{1}{R_{amp1}} + \frac{1}{R_{amp2}} + \frac{2}{r_d} \right)} \\
 \lim_{R_{bias} \rightarrow \infty} \beta &= \frac{\frac{R_{d2}}{2} \left(\frac{1}{R_{d1}} + \frac{1}{R_{d2}} + \frac{2}{r_d} \right) \frac{2}{r_d} + \frac{R_{amp2}}{r_d^2} - \frac{R_{d2}}{r_d^2}}{\frac{R_{d2}}{2} \left(\frac{1}{R_{d1}} + \frac{1}{R_{d2}} + \frac{2}{r_d} \right) \frac{2}{r_d} R_{amp2} \left(\frac{1}{R_{amp1}} + \frac{1}{R_{amp2}} + \frac{2}{r_d} \right) - \frac{R_{amp2}R_{d2}}{r_d^2} \left(\frac{1}{R_{d1}} + \frac{1}{R_{d2}} + \frac{2}{r_d} \right) - \frac{R_{amp2}R_{d2}}{r_d^2} \left(\frac{1}{R_{amp1}} + \frac{1}{R_{amp2}} + \frac{2}{r_d} \right)} \\
 &= \frac{\frac{R_{d2}}{R_{d1}} + 1 + \frac{2R_{d2}}{r_d} + \frac{R_{amp2}}{r_d} - \frac{R_{d2}}{r_d}}{\left(\frac{R_{d2}}{R_{d1}} + 1 + \frac{2R_{d2}}{r_d} \right) \left(\frac{R_{amp2}}{R_{amp1}} + 1 + \frac{2R_{amp2}}{r_d} \right) - \frac{R_{amp2}}{r_d} \left(\frac{R_{d2}}{R_{d1}} + 1 + \frac{2R_{d2}}{r_d} \right) - \frac{R_{d2}}{r_d} \left(\frac{R_{amp2}}{R_{amp1}} + 1 + \frac{2R_{amp2}}{r_d} \right)} \\
 \lim_{R_{bias} \rightarrow \infty} \beta &= \frac{D + \frac{R_{d2}}{r_d} + \frac{R_{amp2}}{r_d}}{\left(D + \frac{2R_{d2}}{r_d} \right) \left(D + \frac{2R_{amp2}}{r_d} \right) - \frac{R_{amp2}}{r_d} \left(D + \frac{2R_{d2}}{r_d} \right) - \frac{R_{d2}}{r_d} \left(D + \frac{2R_{amp2}}{r_d} \right)} = \frac{D + \frac{R_{d2}}{r_d} + \frac{R_{amp2}}{r_d}}{D \left(D + \frac{R_{d2}}{r_d} + \frac{R_{amp2}}{r_d} \right)} = \frac{1}{D} \\
 \frac{u_{OUT}}{i_{sum}} &= \frac{\left[2(R_{d1} || R_{d2}) + \frac{1}{j\omega C_{c1}/2} + r_d + R_{bias} \right] \left[R_{bias} + r_d + \frac{1}{j\omega C_{c1}/2} \right] - R_{bias}^2}{R_{bias}} [\Omega]
 \end{aligned}$$

gain and processing of the smallest signals is required, the diode bridge as a gain controller is disconnected and does not increase the amplifier noise in any way. The penultimate chapter investigates the dependence of the control voltage or current on the gain. Though this dependence is not linear, it does not, however, cause difficulties in autonomous AGC systems. If this dependence would need to be linearized, a functional converter would have to be used. The paper's conclusion was devoted to the analysis of the DC offset of the amplifier and the elimination of some capacitors. The final circuit solution of the amplifier extending its band towards the lower frequencies was presented.

The main novelty of the article is the connection of an operational amplifier with voltage feedback VFB, the gain of which can be continuously changed with the help of one variable resistor while maintaining the bandwidth. The article further highlights the advantages of a diode bridge operating as a variable resistor, especially its low distortion and low noise. The designer will also appreciate the high degree of independence of the amplifier and gain controller.

APPENDIX

Equation (19), as shown at the top of the page.
 Equation (20), as shown at the top of the page.
 Equation (22), as shown at the top of the page.
 Equation (25):

$$A = - \frac{1}{R_{amp1} \left(\frac{1}{R_{amp1}} + \frac{\frac{1}{R_{d2}}}{1 + \frac{r_d}{R_{d1}} + \frac{r_d}{R_{d2}} + \frac{r_d}{R_{bias}} + \frac{r_d^2}{2R_{d1}R_{bias}} + \frac{r_d^2}{2R_{d2}R_{bias}}} \right)} [1]$$

Equation (28), as shown at the top of the page.
 Equation (29), as shown at the top of the page.
 Equation (30), as shown at the top of the page.
 Equation (31), as shown at the top of the page.

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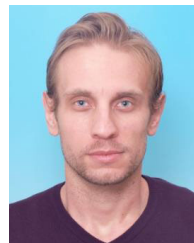
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