SELF-CORRECTING AC AMPLIFIER WITH INDUCTIVE DIVIDER

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Abstract - *In this paper a self-correcting method is presented to reduce the error of an inverting amplifier below 20 ppm in 0-3 kHz frequency range. The method is based on a two step approch. In the first step the phase error of the amplifier is corrected, and then the methode is extended to the amplitude error reduction.*

Keywords - self-correcting inverting amplifier, nonlinear autotransformer modelling, inductive divider

1. INTRODUCTION

The inverting amplifier is a very important building block of an electronic circuit, and widely used in measurement equipments, especially in impedance meter system. There's need for such a precision amplifier, which has an extremely small phase and amlitude error in wide frequency range.

There are some appreciated method at 0-10 kHz frequency range where the accuracy can be improved, but all of them were developed for low voltage level and their usability is focused on this level range. A self-calibrating method is decribed in [1].

In this paper an alternative tecnique is proposed based on self-correcting methode. A suitable divider furthers our goal. It will be shown the theory of self-correcting method, and that, with this method it is possible to reduce the phase error in the range $10^{-6}...10^{-5}$ radian. Then the amlitude error will be examined, and described how will be able to reduce the amplitude error below 20 ppm replacing the resistive divider by an inductive one.

2. THEORY

2.1 Stucture

The inverting amplifier composed of the operational amplifier (A_2), input resistor (R_1) feedback resistor (R_2) is illustrated in Fig. 1 [2,3].The gain of the ideal amplifier is:

$$H = -\frac{R_2}{R_1}. \tag{1}$$

Inasmuch as amplitude and phase error appear due to the finite-amplification, the precision divider (with the elemets R_3, and R_4) generates an error signal V_E, if the ratio of the resistances is:

$$\frac{R_2}{R_1} = \frac{R_4}{R_3} \tag{2}$$

The A_E error signal amplifier forces a compensating current I_C to the input of A_2 and the effect of this second feedback causes a smaller phase and amplitude error than without divider.

In this particular case the A_E error signal amplifier consists of an operational amplifier A_1 and resistor R_5.

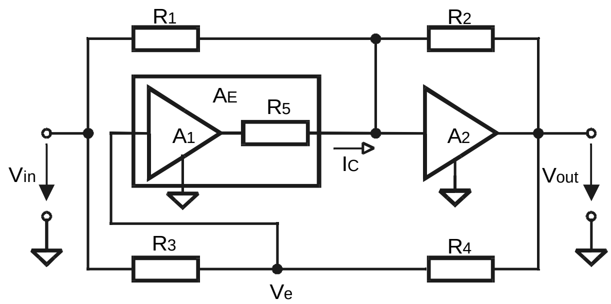

Fig. 1. Block diagram of the self-correcting amplifier

2.2 Analyzing the amplitude and phase error

For the operational amplifiers the following theoretical formula can be written:

$$A_i = \frac{A_{i0}}{1 + j\omega T_i} = \frac{1}{\frac{1}{A_{i0}} + j\frac{\omega}{\omega_{bi}}} \tag{3}$$

where A_{i0} is the gain, T_i is the time constant, and $\omega_{bi} = A_{i0}/T_i$ is the unity-gain circular frequency of A_i amplifier.

The transfer function (4) can be defined from the nodal equasition and the errors can be written according to (4), (5), (6) and the Fig.2. (if $R_i = R$ and gain: -1).

$$F_n(j\omega) = -\frac{A_{20}A_{10} + 2}{A_{20}A_{10} + 2} \cdot \frac{1 + j\omega\frac{2T_1}{2 + A_{10}}}{1 - \omega^2\frac{6T_1T_2}{A_{20}(A_{10} + 2)} + jw\frac{T_1(2A_{20} + 6) + 6T_2}{A_{20}(A_{10} + 2) + 6}}, \tag{4}$$

$$\Delta\varphi(j\omega) = \pi - \left(\operatorname{atan}\frac{2\omega T_1}{A_{10} + 2} - \operatorname{atan}\omega\frac{T_1(2A_{20} + 6) + 6T_2}{A_{20}(A_{10} + 2) + 6 - \omega^2 6T_1T_2}\right) \tag{5}$$

$$e = 1 - \frac{A_{20}A_{10} + 2}{A_{20}A_{10} + 2 + 6}.$$

$$\frac{\sqrt{1 + \left(\omega\frac{2T_1}{A_{10}+2}\right)^2}}{\sqrt{\left(1 - \omega^2\frac{6T_1T_2}{A_{20}(A_{10}+2)+6}\right)^2 + \left(\omega\frac{T_1(2A_{20}+6)+6T_2}{A_{20}(A_{10}+2)+6}\right)^2}} \qquad (6)$$

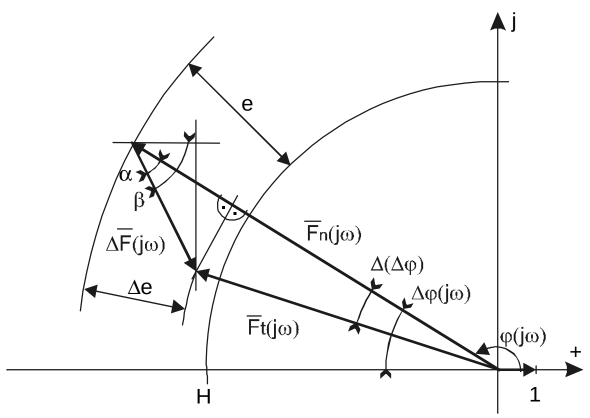

Fig. 2. Phasor diagram of the self-correcting amplifier

The examined amplifier’s amplitude and phase error are responsive to the tolerance of elements, and tolerance-examination is essential for the calculation of error-curves. Supposing that each resistor has a shunt capacitor the condition is wronger. So the composit errors can be defined the sum of nominal error and the tolerance error (according to denotion of Fig.2.):

$$e_{comp} = e_n + \Delta e, \qquad (7)$$

$$\Delta\varphi_{comp} = \Delta\varphi + \Delta(\Delta\varphi). \qquad (8)$$

If the $\Delta\varphi$ is small enough the following relations can be written:

$$\Delta e = \Re(\overline{\Delta F}) \approx \frac{\Delta R_3}{R_3} + \frac{\Delta R_4}{R_4} + \frac{2}{3}\frac{\Delta R_5}{R_5} + \ldots, \qquad (9)$$

$$\Delta(\Delta\varphi) = \Im\overline{\Delta F} \approx 2\frac{\omega}{\omega_1}\left(\frac{\Delta R_1}{R_1} + \frac{\Delta R_2}{R_2}\right) + \ldots. \qquad (10)$$

The change of $\bar{F}$ can be expressed as a total differential.

2.3 Results

Measurements and simulations demonstrate, that if the gain is equal with -1, the phase error can be reduced below 10^{-7} at 50-60 Hz. As the frequency increases the tolerance error has a greater role in (7) and (8). The tolerances of the elements have a different weight in these equations. In (9) the divider elements (R_3, R_4) and their tolerances dominate and the amplitude error can’t be better than 10^{-3} because these two elements set limit of the accuracy. On the other hand the phase error can reach the 1ppm border, because their dominaited elements (R_1, R_2) are weighted by ω/ω_1 factor.

Fig. 3. shows the result of the calculation. As can be seen, the phase error more than three orders of magnitude smaller than without resistance ladder. As can be seen also that the amplitude error do not reach the desired 10^{-5} border.

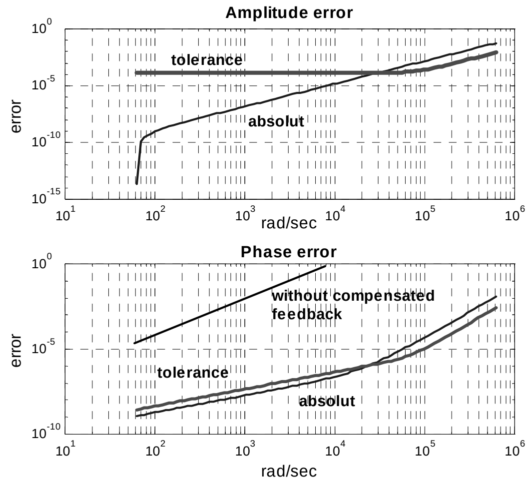

Fig. 3. Result the error calculation

3. MODIFIYING THE DIVIDER

3.1 Equivalent network

In Fig.4. the modificated circuit block-diagram can be seen. The resistance divider is exchanged for an inductive one, and the number of turns ratio (N_1, N_2) definite the V_E error voltage. It’s a real anticipation that the amplitude error will be smaller by the the help of inductive divider without increasing the phase error.

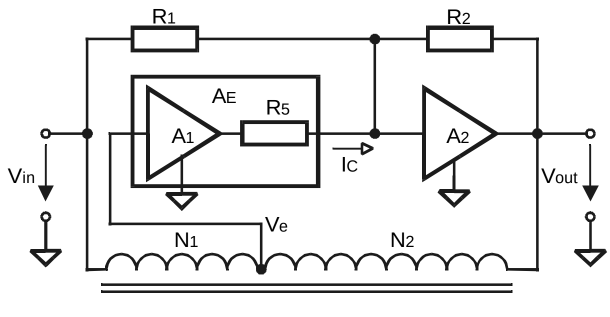

Fig. 4. Modificated circuit block-diagram

The high accuracy requirements reawekened the physical model that describe the operation of inductive divider as a function of frequency, with various leakage and stray parameters also taken into account. In Fig. 5. the detailed circuit diagram can be seen. The resistors (R_1, R_2, R_5) are shown as adminttances, because during the calculation of errors all this kind of elements are modelled as a resistor with a parallel C capacitance ($Y_i = 1/R_i + j\omega C_i$). The equivalent network components of the inductive divider are the main field inductance (L_{MF}), iron losses (R_{IL}), stray capacitance (C_S), leakage inductances (l_3, l_4) and copper resistances

(r_3, r_4). L_{MF} and R_{IL} are nonlinear elements, their values are dependent on the frequency and voltage level. The two elements can be modelled in the range where the effect of stray capacitance is still absent. The estimation of the stray capacitance is written in [3]. In this case the divider is in no-load mode so the secondary elements are negligible, and the primary side elements are distributed according to the windings N_1 (l_3 and r_3), N_2 (l_4 and r_4), and the inputs of the ideal divider D are connected between V_1 and V_2 potentials. The output voltage of the inductive divider is supplied by V_1 and V_2 according to the picture in Fig. 5. and the following equation:

$$V_e = \frac{V_{out}N_1 + V_{in}N_2}{N_1 + N_2} \tag{11}$$

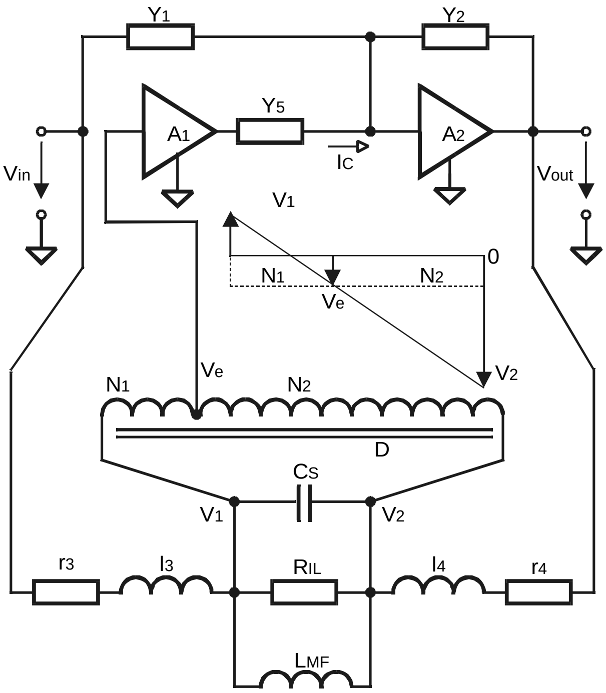

Fig. 5. Block diagram with equivalent network of the inductive divider

3.2 Results

Here also the tolerance error has a greater part. It'is necessary to find which part of this implementation is the most sensitive to this kind of error, and how the value of these elements affect it. In the Fig. 6. and in the equations (12)...(19) only those errors ($\delta F/\delta X_i \cdot \Delta X_i$) are calculated and plotted against frequency which have significant rule, so the crossbranch element of the inductive divider and the parallel capacitances are absent.

Disregarding the derivation and complating the shotening the following equation can be given:

$$\frac{\delta F}{\delta R_1} \approx (1+K)\frac{R_5}{R_1^2} \cdot \left[\frac{1}{A_{10}} + \frac{j\omega}{\omega_{11}}\right], \tag{12}$$

$$\frac{\delta F}{\delta R_2} \approx -K \cdot (1+K)\frac{R_5}{R_1^2} \cdot \left[\frac{1}{A_{10}} + \frac{j\omega}{\omega_{11}}\right], \tag{13}$$

$$\frac{\delta F}{\delta \omega_{11}} \approx -\frac{j\omega}{\omega_{11}^2}K(1+K)\left[1 + \frac{R_5}{R_1}\right]\left[\frac{1}{A_{20}} + \frac{j\omega}{\omega_{12}}\right], \tag{14}$$

$$\frac{\delta F}{\delta \omega_{12}} \approx -\frac{j\omega}{\omega_{12}^2}K(1+K)\left[1 + \frac{R_5}{R_1} + \frac{R_5}{R_2}\right]\left[\frac{1}{A_{10}} + \frac{j\omega}{\omega_{11}}\right], \tag{15}$$

$$\frac{\delta F}{\delta r_3} \approx jK^2 \cdot \left[\frac{j}{R_{IL}} + \frac{1}{\omega L_{MF}} - \omega C_S\right], \tag{16}$$

$$\frac{\delta F}{\delta r_4} \approx j(1+K) \cdot \left[\frac{j}{R_{IL}} + \frac{1}{\omega L_{MF}} - \omega C_S\right], \tag{17}$$

$$\frac{\delta F}{\delta l_3} \approx K^2 \cdot \left[\frac{j\omega}{R_{IL}} + \frac{1}{L_{MF}} - \omega^2 C_S\right], \tag{18}$$

$$\frac{\delta F}{\delta l_4} \approx -(1+K) \cdot \left[\frac{j\omega}{R_{IL}} + \frac{1}{L_{MF}} - \omega^2 C_S\right], \tag{19}$$

where $K = N_2/N_1$.

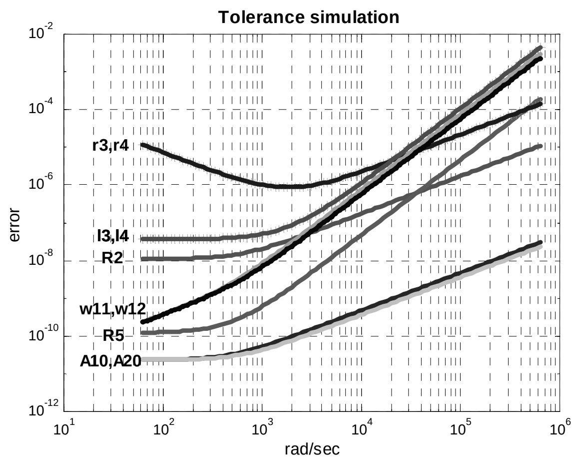

Fig. 6. Tolerance simulation ($H = 10$)

4. CONCLUSION

From the equation can be derived the following statements:

- if the ratio of the windings increases the most important error components increases also, and at the given frequency there's a chance to minimize the these components' values. It's necessary to turn especial attention to the fabrication of the inductive divider, (16-19) and to set the point to the work frequency.

- There's no need a special error amplifier (A_1), because it has no ability to decrease these value in tems (16-19)

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