Phase control as basis for precise impedance bridge microminiaturization

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Abstract

The new approach to the creation of the equivalent voltage dividers is proposed. This approach is based on the synthesis of the signal, having digitally controlled magnitude by the algebraic summing of the signals, having digitally controlled phase. On this base one of the possible bridges structure has been developed and its properties have been analyzed

Key words: phase control, precise impedance parameters measurements.

Introduction

For the precise impedance measurements on audio frequency range in main laboratories usually simple [1-13] or quadrature [14-21] transformer bridges are used. Such bridges contain, as main part, some (2-6 or more) precision transformer dividers. It led to great dimensions of the devices, their narrow frequency range and high cost. These disadvantages sharply increase on the lower frequencies. It makes impossible creation of the transformer bridge for low frequencies. Therefore, replacing of the inductive divider in impedance measurements by another device could be very useful.

Last time the digital synthesis of the signals and creation on this basis of different bridges are widely used [22-25]. But these bridges can't compete in accuracy with transformer bridges.

So, the main today problem consist in the following: most accurate transformer bridges have great dimension and cost and really can't operate on low frequency range.

Solution

The main idea of this article is shown on Fig .1. Here, on the plane of complex numbers (1; j.), are shown basic signal (vector U_0), which coincides with real axis and two additional vectors U_{11} and U_{21}.

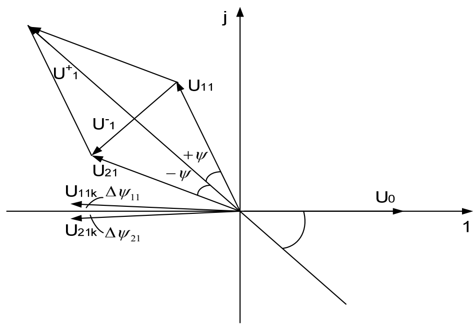

Fig.1

The bisectrix, dividing onto two angle between vectors U_{11} and U_{21}, is turned relatively on the real axis on angle φ.

Basic and additional signals (vectors U_0, U_{11} and U_{21}) could have different magnitudes and phases but it is preferable to use the signals, satisfying to the equality:

$$|U_0| = |U_{11}| = |U_{21}| \; ; \; \psi_{11} = -\psi_{21}. \qquad (1)$$

We will sum additional vectors U_{11} and U_{21}. The total balancing vector U_1^+ will lie on the bisectrix. It is easy to show that this signal U_1^+ is described by equation:

$$U_1^+ = 2|U_0|\cos\psi \sin(\omega t + \varphi). \qquad (2a)$$

Let us subtract additional vectors U_{11} and U_{21}. (the subtraction can be submitted by summing of two signals one of them being negative). The differential balancing vector U_1^- will be perpendicular to the bisectrix. It is easy to show that the signal U_1^- is described by equation:

$$U_1^- = 2|U_0|\sin\psi \cos(\omega t + \varphi). \qquad (2b)$$

We will ***change*** angles ψ and –ψ simultaneously from zero to the same current value $|\psi|$.

Equations (2a) and (2b) evidently show that balancing signal U_1 in both these cases could be changed in magnitude and phase **by controlling of the phases ψ and φ only**.

The range ***changing*** of the U_1 magnitude in both cases lie between 0 and $\pm 2U_0$. The range ***changing*** of its phase lies between 0 and ± 180. It is obviously that using the signals U_0 and U_1 we could create universal balanced bridges for measurements of any type impedances using the ***changing*** of the phase only.

The dependence of the vector U_1 magnitude on the phase angle ψ is nonlinear. It doesn't influence on the accuracy of the measurement because of this dependence is strictly calculated. But the slope of this dependence (see equations (2a) and (2b)) change in the whole range of angle ψ from 0 to 1. The zero value of the slope could create difficulties in automated bridge balancing. If we will limit the range of the vector U_1 changing, for example, to $|U_1| \le \sqrt{2}|U_0|$ the slope will change from 0.5 to 1 only. Such change of the slope will not influence the balancing process and only slightly ($\sqrt{2}$) restricts range of measurement.

Described above system of vectors isn't unique, which permits to create the balancing vector, controlled in magnitude by phase control. It is just the simplest one only. The calibration procedure of the bridges, using this system, is the simplest as well.

Signals U_0, U_{11} and U_{21} could be easily created using today digital technique. For this purpose usually the different synthesis circuit (synthesizers), based on the fast digital-to-analog converters, are used [7, 8.].

Such synthesizers in audio frequency range have rather great uncertainty - usually around 10^{-4} or worse. But these synthesizers could have enough good short term (during one measurement - one minute or less) stability.

If the initial vectors magnitude ratios and the phase sift differences from the nominal would be determined and excluded, only its short term instability and accuracy of the phase ***changing*** will influence on the common uncertainty of the measurement.

Using proper components, their temperature stabilization, etc. synthesizers short term instability could be reduced to values 10^{-8} or less.

Phase **changing** uncertainty is restricted by components phase noise only. Today this value could be easily reduced to 10^{-9} or less.

Of course, the components noise, caused by different sources (low frequency noise, noise, caused by charge injection in DAC, etc) will influence the measurement resolution.

The lot of common problems, arising during the impedance measurements and widely discussed and solved in lot of excellent classic works, (such as four pair terminal connections, etc) could restrict the measurement uncertainty as well. But we will not go here into all these very interesting and important problems.

A lot of different bridges could be created using described approach. Here we will show the simplest bridge with current comparison and on this example will discuss the peculiarities of its balancing and problems of its calibration.

Fig.2 shows the functional diagram of the possible bridge, using proposed idea. The bridge consists of the stable DC voltage source $U_=$, connected to inputs of three synthesizers S_{11}, S_{21} and S_0. These synthesizers create appropriate sinusoidal signals U_{11}, U_{21} and U_0 under the microcontroller MC control.

The output sinusoidal signals of the synthesizers S_{11}, and S_{21} are summed by adder Σ, creating balancing signal U_1. The impedances to be compared Z_0 and Z_x are connected to the outputs of synthesizer S_0 and adder Σ. The vector voltmeter VV measures the bridge output signals through switcher C_2. The microcontroller MC processes the results of the vector voltmeter VV measurement and control all the bridge balancing procedure by changing of the appropriate signal phases.

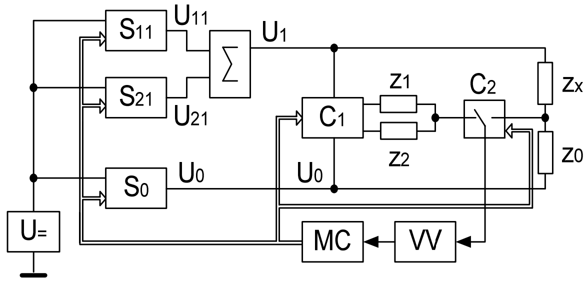

Fig.2

Bridge balancing.

The balance condition of the bridge is described by very simple equation:

$$\frac{Z_x}{Z_0} = \frac{\vec{U}_{1b}}{\vec{U}_{ob}} \qquad (3)$$

Let we'll balance the bridge, changing the magnitude and the phase φ of the vector U_1. In this case the equation could be rewritten in the form:

$$\frac{Z_x}{Z_0} = \frac{\vec{U}_{1b}}{\vec{U}_{ob}} = (2\cos\psi_b)e^{-j\varphi_b} = \rho_b e^{-j\varphi_b} \quad (4)$$

Equation (4) shows with evidence that we can get direct reading, if we will measure the magnitude and phase of the ratio $\frac{Z_x}{Z_0} = \rho_x e^{-j\varphi_x}$,

where: $\left|\frac{Z_x}{Z_0}\right| = \rho_x$.

In this case the balancing equation (4) divides onto two simplest:

$$\rho_x = \rho_b = 2\cos\psi_b \quad \text{and} \quad \varphi_x = \varphi_b . \qquad (5)$$

where: ψ_b and φ_b – the coordinates of balance point.

Of course, if we know these two parameters of impedance ratio any other desired parameters of this ratio, as well as the impedance Z_x parameters, can be easily calculated.

To balance the bridge we could use variational method [10]. In this case we firstly measure initial bridge unbalance signal U_{n1}. After it we provide the variation of the bridge parameters, used for its balancing (the angles ψ or φ), and measure new unbalance signal. For certainty, let we will change the ψ, adding $\Delta\psi_v$ and measure the new unbalance signal U_{n2}. Next system of equations will describe this process:

$$\rho_x e^{-j\varphi_x}|U_0| - |U_0|\rho e^{-j\varphi} = \vec{U}_{n1}(Z_x + Z_0)/Z_0$$
$$\rho_x e^{-j\varphi_x}|U_0| = |U_0|(\rho + \Delta\rho_v)e^{-j\varphi} = \vec{U}_{n2}(Z_x + Z_0)/Z_0 \quad (6)$$

Here $\psi = \psi_b + \Delta\psi$ and $\varphi = \varphi_b + \Delta\varphi$, where: $\Delta\psi$ and $\Delta\varphi$ is the distance between coordinates of current bridge point (ψ, φ) and coordinates (ψ_b, φ_b) of point of balance. Solving the system (6) we find these distances from next implicit function:

$$\frac{\rho_x}{\rho} e^{-j\Delta\varphi} = (1 - A\delta_v); \quad (7)$$

where:

$$A = \frac{\vec{U}_{n1}}{\vec{U}_{n2} - \vec{U}_{n1}} = \left|\frac{\vec{U}_{n1}}{\vec{U}_{n2} - \vec{U}_{n1}}\right| e^{-j\varphi_a} = |A| e^{-j\varphi_a};$$

$$\delta_v = \frac{\cos(\psi + \Delta\psi_v)}{\cos\psi} - 1.$$

Calculating and entering in the bridge the coordinates of balance point, we achieve the full bridge balancing.

Bridge calibration

To calibrate the bridge let we will perform balance equation. It is easy to see that additional signals $\vec{U}_{11}$ and $\vec{U}_{21}$ could be described by equations: $\vec{U}_{11} = \vec{U}_{11n} + \Delta\vec{U}_{11} = \vec{U}_{11}(1 + \delta_{11})$ and $\vec{U}_{21} = \vec{U}_{21n} + \Delta\vec{U}_{21} = \vec{U}_{21}(1 + \delta_{21})$. These signals have constant nominal values $\vec{U}_{11n}$ and $\vec{U}_{21n}$ and constant relative deviations from nominal δ_{11} and δ_{21} as well. Using these formulas we could rewrite balance equation into two equivalent forms:

$$\frac{Z_x}{Z_0} = \frac{\vec{U}_{11n} + \Delta\vec{U}_{11}}{\vec{U}_0} + \frac{\vec{U}_{21n} + \Delta\vec{U}_{21}}{\vec{U}_0} = = \frac{\vec{U}_{11n}}{\vec{U}_0}(1 + \delta_{11}) + \frac{U_{21n}}{U_0}(1 + \delta_{21}). \quad (8a)$$

or:

$$\frac{Z_x}{Z_0} = \frac{\vec{U}_{11n} + \vec{U}_{21n}}{\vec{U}_0} + \frac{\Delta\vec{U}_{11} + \Delta\vec{U}_{21}}{\vec{U}_0} = = \frac{U_{1n}}{U_0} + (\delta_{11} + \delta_{21}). \quad (8b)$$

where: $\delta_{11} = \Delta U_{11}/U_0$ and $\delta_{21} = \Delta U_{21}/U_0$ - relative deviations of the additional vectors U_{11} and U_{21} from nominal values U_{11n} and U_{21n}. Values δ_{11} and δ_{21} don't depend on the angles ψ and φ.

Equation forms (8a) and (8b) correspond to two different calibration procedures. Both these procedures are based on consistent nulling of one of the components in formulas (8a), (8b) and determination of the remains component.

First calibration procedure, based on formula (8a), consists of two similar steps:

1. Calibration of the synthesizers S_{11}.
2. Calibration of the synthesizers S_{21}.

1. On the first step of the calibration procedure we switch off the operation of the synthesizer S_{21} (for example, entering and maintaining zero control codes into the synthesizer S_{21} or simply switching its output from the appropriate adder input).After it we set into synthesizer S_{11} the codes, corresponding to equality U_0= -U_{11n}. In this case the outputs of the synthesizer S_0 and adder could be considered as input and output of the appropriate inverter and the calibration procedure, described in [9], could be used.

Calibration circuit, consisting of standards Z_1 and Z_2, is connected to outputs of the synthesizer S_0 and adder Σ through switcher C_1. Last one reverses phase of the calibration circuit connection to the mentioned signal sources during the calibration process (fig.2).

All procedure is based on variation and replacing methods [26, 27] and consists of three stages.

1. At the first stage, the switcher C_2 connects the vector voltmeter to the output of the Z_1-Z_2 divider. Switcher C_1 remains in the initial position and the vector voltmeter measures the unbalance signal $\vec{U}_{n1}$.
2. At the second stage, the MC varies the synthesizer S_{11} transfer coefficient on δ_{11v}. After that, thevoltmeter measures the unbalance signal $\vec{U}_{n2}$.
3. At the third stage, the MC reverses the switcher C_2 and the vector voltmeter measures the signal $\vec{U}_{n3}$.These measurements are described by the following system of equations:

$$\vec{U}_0 - \frac{\vec{U}_0 - \vec{U}_0(1 + \delta_{11})}{Z_1 + Z_2} \cdot Z_1 - \vec{U}_{n1} = 0,$$
$$\vec{U}_0 - \frac{\vec{U}_0 - \vec{U}_0(1 + \delta_{11} + \delta_{11v})}{Z_1 + Z_2} \cdot Z_1 - \vec{U}_{n2} = 0, \quad (9)$$
$$\vec{U}_0 - \frac{\vec{U}_0 - \vec{U}_0(1 + \delta_{11})}{Z_1 + Z_2} \cdot Z_2 - \vec{U}_{n3} = 0.$$

Neglecting the second order terms we will get the next result:

$$\delta_{11} = \frac{\delta_{11v}}{2} \cdot \frac{\vec{U}_{n1} - \vec{U}_{n3}}{\vec{U}_{n2} - \vec{U}_{n1}}. \quad (10)$$

So, using the results of the three mentioned measurements and equation (10), we can find the synthesizer S_{11} relative deviation of the transfer coefficient from nominal. The result of the measurement of the deviation does not depend on additive or multiplicative voltmeter errors. The accuracy of the error synthesizers S_{11} determination depends on the voltmeter nonlinearity and sensitivity only.

2. On the second step of the calibration procedure we switch off the operation of the synthesizer S_{11}. After it we set into synthesizer S_{21} the codes, corresponding to equality $U_0 = -U_{21n}$ and repeat described previously calibration procedure, including next three measurements. In such way we get the value $\delta_{21} = \Delta U_{21}/U_0$.

It is easy to show that to correct the result of the ratio Z_x/Z_0 measurement, the U_1 has to be divided on $1 + \delta_{11} + \delta_{21}$.

Described calibration procedure doesn't needs using of accurate standards in calibration circuit. It needs their short term stability during the calibration process only. But it needs six measurements, which could take rather long time. Using of the relatively high resistive calibration divider decrease the possible power of the signals to be measured as well.

Another calibrating procedure, taking less time and less number of measurements, is illustrated by Fig.3. This procedure is based on second form of balance equation (8b).

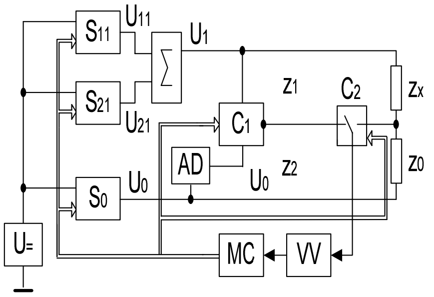

Fig.3.

Here the additional divider AD is connected to the output of the synthesizer S_0. This divider transfers the synthesizer S_0 output signal with transfer coefficient δ_v. The AD and Adder outputs are connected to the switch C_1 inputs. Calibration procedure consists of two measurements only.

1. During first step we set into synthesizers S_{11} and S_{21} the codes, which correspond to equality $U_{11n} = -U_{21n}$. In this case on the adder output will act the signal $\Delta U_{11} + \Delta U_{21}$. Through switchers C_1 and C_2 the vector voltmeter measures this signal as U_{n1}.

2. On the second step the switcher C_1 is turned of, so that vector voltmeter measure the AD output signal U_{n2}. These two measurements are described by simple system:

$$\begin{aligned} \Delta U_{11} + \Delta U_{21} &= U_{n1} \\ \delta_v U_0 &= U_{n2} \end{aligned} \qquad (11)$$

It is easy to show that the total deflection from nominal here will be determined by equation:

$$\delta_{11} + \delta_{21} = \frac{U_{n1}}{U_{n2}} \delta_v \qquad (12)$$

As earlier, this result can be used for uncertainty correction.

Uncertainty of the value $\delta_{11} + \delta_{21}$ estimation doesn't depend on vector voltmeter gain error. It depends, as earlier, on its nonlinearity and sensitivity. In addition, it depend on vector voltmeter zero shift. This one has to be measured and eliminated from results of measurements using well known procedures. In case of the second calibration procedure correction uncertainty depends on additive divider AD accuracy as well. For example, if the $\delta_{11} + \delta_{21} = 10^{-4}$ and we want to get component of common uncertainty, created by $\delta_{11} + \delta_{21}$, being less than 10^{-8}, the uncertainty of AD transfer coefficient δ_v has to be less than 10^{-4}.

Preliminary experimental investigations of the described bridge have shown that uncertainty of the measurement in real condition, without temperature stabilization of the used components, achieve 10^{-7}. These investigations have shown as well, that bridge resolution depends on the noise, caused by charge injection in used DAC. This noise quickly increase then operation frequency increase. On low and infralow frequency range the contribution of this noise to common uncertainty sharply decrease. But on this frequency range the flicker noise increase its influence on the result of measurement.

Conclusion

Two features of this bridge have to be underlined:

Firstly, we have got possibility to measure accurately impedances in wide and continuous range of values without using of the inductive dividers.

And, second, we have got possibility to measure impedances accurately in wide frequency range.

In principle, there are no any limitations in region of low frequency (the time of measurement only).

In high frequency range there is the limitation, caused by operation speed and number of digits in synthesizers. Last one determines the discreteness of bridge balancing only. Regard to the fact that today digital-to-analog converters have the operating frequency more than 10 MHz and discreteness 10^{-3}-10^{-5}, we could create bridges with highest operating frequency up to units of kHz.

Preliminary experimental investigations have shown that theoretical ideas of this article could be successfully realized and we could create on this basis bridges, measuring impedance ratio with uncertainty up to 10^{-7} and less.

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