Quadrature method for broadband measurement of microwave path parameters

Yu Gimpilevich and Yu Tyschuk
Sevastopol State University, 33, Universitetskaya St., Sevastopol, 299053, Crimea, Russian Federation

E-mail: y.tyschuk@gmail.com

Abstract. A mathematical model has been developed for the procedure for broadband automatic measurement of the complex reflection coefficient and power level in the microwave path based on the quadrature measurement method. Analytical expressions are obtained for the quadrature components at the outputs of a quadrature demodulator connected with a switch to two non-directional measuring probes for the case of an unmatched microwave path. A modified system of nonlinear equations with respect to measured values was formed and solved, which allows setting zero assumptions. A block diagram of a broadband device for built-in automatic control of microwave channel parameters has been developed, which implements the developed measurement procedure and the solution of a system of measurement equations to determine the current values of the power level factor, as well as the modulus and argument of the complex reflection coefficient.

1. Introduction
In modern microwave technology, there are a number of methods for measuring complex parameters of microwave paths, based on the branching of signals from these paths using non-directional probes. In view of the relative simplicity of the hardware implementation, it is on the basis of these methods that devices for built-in monitoring of microwave path parameters are built.

In works [1 - 4], methods of measurements based on quadratic amplitude detection of branched signals from various cross-sections of microwave paths are considered, which allows to restore the amplitude distribution of the electromagnetic wave in the transmission line, and from it to calculate the modulus and argument of the complex reflection coefficient. Devices based on these methods have insufficient sensitivity and low measurement accuracy due to the small dynamic range of square-law amplitude detectors.

In [5 - 7], a method for measuring the modulus and argument of the complex reflection coefficient was developed, based on the application of the principle of quadrature detection of branched signals. The quadrature method significantly expands the dynamic range (not less than 30 dB) and allows measurements based on both the amplitude and phase distribution of the field in the transmission line. The disadvantage of this method is that the mathematical model of the measuring procedure [5, 6] is developed for one frequency of the probing signal, which leads to a narrow band of built-in control devices [8, 9].

A new mathematical model of the procedure for automatic measurement of the complex reflection coefficient and the power level in the microwave path is developed below, eliminating this drawback.
In this work, analytical expressions are obtained for the quadrature components at the outputs of a quadrature demodulator, alternately connected to the outputs of the measuring probes, for the case of an arbitrary value of the frequency of the probing signal, which makes it possible to eliminate the frequency measurement error and, thereby, to significantly increase the broadband of the built-in device.

2. Block diagram of the device that implements the quadrature method of broadband measurement of the microwave path parameters

The block diagram of the device is shown in figure 1. The device consists of 2 blocks: microwave sensor; information processing unit (IPU). The IPU includes an analogue-to-digital processing (ADP) module and a display module (DM).

The microwave sensor includes a directional incident coupler (DC) and two fixed non-directional probes (P1 and P2). The measured load ML and the microwave generator (MG) are connected to the microwave sensor through the standard C1 and C2 connectors, respectively.

The distances between the load connection plane ML and the connection planes of the first P1 probe and the second P2 probe in figure 1 are labelled as $l_1$ and $l_2$ respectively. At the output of the incident wave of the directional DC coupler, a microwave oscillation $u_0(t)$ with a constant amplitude is formed, proportional to the amplitude of the incident wave in the path, which is further used as a reference signal. At the outputs of the non-directional probes P1 and P2, two microwave oscillations $u_1(t)$ and $u_2(t)$ are formed, which are then used as measuring signals in the quadrature detection circuit. The signals $u_0(t)$, $u_1(t)$, $u_2(t)$ from the microwave sensor unit go further to the information processing unit (IPU).

In the IPU, the signal $u_0(t)$ is fed to the reference input of the QD quadrature demodulator, and the measurement signals $u_1(t)$ and $u_2(t)$ are fed to the 1st and 2nd inputs of a three-channel microwave switch (S), to the 3rd input of which a matched load (MtL) with a 50 Ohm resistance is connected. The microwave switch S, according to the commands coming from the MC microcontroller, alternately connects the three inputs of this switch to the measuring input of the QD quadrature demodulator. As a result of quadrature detection in the first two switching cycles, two pairs of quadrature components appear sequentially in the spectra of the output voltages QD: $I_1$, $Q_1$ (when the first P1 probe is connected) and $I_2$, $Q_2$ (when the second P2 probe is connected). These components carry information about the measured parameters of the load ML. In the third cycle, the measurement input QD is potential-free, since it is connected to the housing via the terminated MtL load. The latter circumstance leads to the fact that an offset voltage (zero drift) is formed at the QD output, which is subtracted from the quadrature components in MC, which eliminates the influence of zero drift on the measurement results. Low-pass filters LPF1 and LPF2 are used for the quadrature components comprehension. The voltages from these filters are fed to an analogue-to-digital converter (ADC), where they are amplified and converted into digital code. From the ADC output, digital signals are fed to the MC microcontroller. MC carries out control of the microwave switch S, primary processing of measurement information (averaging of measurement results of quadrature components, elimination of zero drift), solution of the system of measurement equations and delivery of measurement results to the display.
Thus, as a result of optimization of the device structure, the following was achieved:

- the distance between the probes in the microwave sensor of the optimized device can be any, which allows to create a broadband device by developing new measurement and calibration algorithms;
- in the process measurements, only one channel of quadrature detection is used, which makes it possible to increase the accuracy, since it eliminates the error due to the dispersion of the characteristics of the quadrature demodulators;
- the influence of zero drift on the measurement result is practically excluded by periodically correcting the values of the quadrature components, which also leads to an additional increase in the measurement accuracy;
- all mathematical calculations are carried out in the MC, which eliminates the need to use a personal computer to process the measurement results.

Figure 1. Block diagram of a broadband quadrature meter of microwave path parameters.
3. Development of a mathematical model of the measurement procedure

We will develop a mathematical model under the assumption that the elements of the measurement circuit are ideal (C1, C2, DC, QD, etc.) and the absence of noise.

We will assume that the measured load ML is connected to the microwave sensor through connector C1. In this case, the complex reflection coefficient (CRC) of the measured load in the plane of its connection will be written in the form

$$\Gamma = |\Gamma|e^{j\varphi},$$

(1)

where $|\Gamma|$ and $\varphi$ are the module and the CRC argument to be measured.

As a result of the interference of the incident and reflected waves in the transmission line, a mixed wave regime occurs.

The amplitude $|E_i|$ and the initial phase $\phi_i$ of the total wave in the plane of the P1 probe connection, taking into account (1), can be written by the following expressions:

$$|E_i| = E_n\sqrt{1 + |\Gamma|^2 + 2|\Gamma|\cos(\varphi - 2\beta l_i)}.$$

(2)

$$\phi_i = \beta l_i + \arctg\left[\frac{|\Gamma|\sin(\varphi - 2\beta l_i)}{1 + |\Gamma|\cos(\varphi - 2\beta l_i)}\right].$$

(3)

The expression for the harmonic oscillation branched off by the P1 probe and fed to the measurement input of the QD quadrature demodulator in the first measurement cycle using (2) and (3) can be written as follows:

$$u_i(t) = k_{n1}|E_i|\cos(\omega_0 t + \phi_i + \theta_1),$$

(4)

where $k_{n1}$, $\theta_1$ — coefficient of transmission and phase shift of the first branch channel to the measuring input QD, respectively; $\omega_0$ — the circular frequency of the microwave oscillation; $t$ — current time.

Similarly, you can write the expression for the harmonic oscillation, branched off by the P2 probe and fed to the measuring input of the QD quadrature demodulator in the second measurement cycle:

$$u_2(t) = k_{n2}|E_2|\cos(\omega_0 t + \phi_2 + \theta_2),$$

(5)

where $k_{n2}$, $\theta_2$ — the gain and phase shift of the second branch channel to the measurement input QD, respectively.

Let us write the expressions for the reference oscillation $u_o(t)$, which is formed in the microwave sensor using DC:

$$u_o(t) = k_o E_n \cos(\omega_0 t + \psi_o),$$

(6)

where $k_o$, $\psi_o$ is the transmission coefficient and the phase shift of the reference signal generation channel to the reference input QD.

In the QD quadrature demodulator, two-channel multiplication of the measurement and reference oscillations is carried out, and in the second channel the reference signal is pre-shifted in phase by $-90^0$ and limited in amplitude. Taking into account formulas (4) - (6), we write down the expressions for these products when the oscillation $u_i(t)$, branched off by the P1 probe is fed to the measuring input QD:
\[
    u_i(t) \cdot u_o(t) = \frac{1}{2} k_i U_0 \left[ \cos(\phi_i + \theta_i - \psi_o) + \cos(2\omega_c t + \psi_o + \phi_i + \theta_i) \right]; \tag{7}
\]
\[
    u_i(t) \cdot \overline{u_o(t)} = \frac{1}{2} k_i U_0 \left[ \sin(\phi_i + \theta_i - \psi_o) + \sin(2\omega_c t + \psi_o + \phi_i + \theta_i) \right]. \tag{8}
\]

where \( u_i(t) \) the reference signal \( u_o(t) \) shifted by \(-90^\circ\); \( k_i = k_{u1}k_{o} \); \( U_0 = \text{const} \) — the amplitude of the reference signal after limiting.

Thus, constant (slowly varying) \( I_1 \) and \( Q_1 \) components appear in the spectra of the output voltages of the QD quadrature demodulator, and the components are the first terms in expressions (7) and (8). The spectra also contain components whose frequencies are twice the frequency of the microwave oscillation - these are the second terms in expressions (7) and (8). The signals from the QD outputs go to the inputs of the low-pass filters LPF1 and LPF2, which provides the selection of low-frequency \( I_1 \) and \( Q_1 \) components and suppression of high-frequency components.

For convenience, we introduce the concept of the relative level of the incident wave amplitude in the path \( K_E \) — this is a dimensionless quantity directly proportional to the incident wave \( |E_n| \) amplitude. At the nominal power level in the path on which the instrument is calibrated, it is advisable to set the value \( K_E \) equal to unity.

It can be shown that all the required parameters are related to four measured quadrature components by the following system of measuring equations:

\[
\begin{align*}
    I_1 &= K_{i1} K_E \sqrt{1 + X^2 + Y^2 + 2(X \cos(\phi_{01}) + Y \sin(\phi_{01}))} \cdot \cos \left( \theta_{01} + \arctg \frac{Y \cos(\phi_{01}) - X \sin(\phi_{01})}{1 + X \cos(\phi_{01}) + Y \sin(\phi_{01})} \right); \\
    Q_1 &= K_{i0} K_E \sqrt{1 + X^2 + Y^2 + 2(X \cos(\phi_{01}) + Y \sin(\phi_{01}))} \cdot \sin \left( \theta_{01} + \arctg \frac{Y \cos(\phi_{01}) - X \sin(\phi_{01})}{1 + X \cos(\phi_{01}) + Y \sin(\phi_{01})} \right); \\
    I_2 &= K_{i0} K_E \sqrt{1 + X^2 + Y^2 + 2(X \cos(\phi_{02}) + Y \sin(\phi_{02}))} \cdot \cos \left( \theta_{02} + \arctg \frac{Y \cos(\phi_{02}) - X \sin(\phi_{02})}{1 + (X \cos(\phi_{02}) + Y \sin(\phi_{02}))} \right); \\
    Q_2 &= K_{i0} K_E \sqrt{1 + X^2 + Y^2 + 2(X \cos(\phi_{02}) + Y \sin(\phi_{02}))} \cdot \sin \left( \theta_{02} + \arctg \frac{Y \cos(\phi_{02}) - X \sin(\phi_{02})}{1 + (X \cos(\phi_{02}) + Y \sin(\phi_{02}))} \right),
\end{align*}
\]

where \( X = |\hat{I}| \cos \phi \); \( Y = |\hat{I}| \sin \phi \).

Computer simulation of the solution of the system of equations (9) using an iterative procedure confirmed the possibility of solving this system with zero initial approximations (initial guesses).

After solving the system of equations with respect to unknowns \( X, Y, K_E \) the modulus and argument of the CRC are calculated by the formulas:

\[
    |\hat{I}| = \sqrt{X^2 + Y^2}; \tag{10}
\]
\[
    \phi = \arctg \frac{Y}{X}. \tag{11}
\]

The system of equations (9) includes constants \( K_{i0}, K_{i1}, K_{i0}, K_{i0}, \theta_{01}, \theta_{02}, \varphi_{01}, \varphi_{02} \), which are determined at the stage of the instrument calibration procedure.
4. Conclusions
A mathematical model of a wide-band quadrature measurement method has been developed, as a result of which a system of four measuring equations is obtained, valid for any frequency of the operating range of the built-in device. This system of equations connects the signals at the outputs of the quadrature demodulator with the measured parameters.

The modelling of the solution of the system of measuring equations in the frequency range equal to one octave is carried out and it is shown that the solution is achieved at zero initial approximations (initial guesses) for the sought variables.

A block diagram of an automatic broadband built-in measuring device that implements the proposed algorithm has been developed. It is shown that the device produces two pairs of quadrature signals $I_1$, $Q_1$, $I_2$, $Q_2$, which make it possible, by solving the measuring equations, to determine the modulus and argument of the complex reflection coefficient, as well as the signal power level in the transmitting channel.

A solution is proposed that allows apply only one quadrature demodulator and virtually eliminate the influence of zero drift on the results, which together improves the measurement accuracy.

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References
[1] Abubakirov B A, Gudkov K P and Nechaev E V 1984 Measurement of Parameters of Radio Engineering Circuits (Moscow: Radio and communication)
[2] Gimpilevich Yu B and Smailov Yu Y 2005 A Method for measuring of two microwave signals vector ratio Proc. of the 5th IEEE Int. Conf. on Antenna Theory and Techniques 397-98
[3] Gimpilevich Yu B and Noskovich V I 2007 Calibrated complex reflectance meter on the basis of two-channel microwave transducer Telecommunications and Radio Engineering 66(4) 363-71
[4] Gimpilevich Yu B 2008 The estimation of built-in devices measurement error influence on the authenticity of microwave units parameters monitoring Proc. Int. Conf. on Modern Problems of Radio Engineering, Telecommunications and Computer Science, TCSET-2008 (Lviv—Slavsko, Ukraine) 518-19
[5] Gimpilevich Yu B and Zebek S E 2014 The mathematical model of a complex reflection coefficient measuring instrument based on a method of direct frequency conversion Proc. of the 24th IEEE Int. Conf. on Microwave & Telecommunication Technology, CriMiCo-2014 (Sevastopol, Crimea) 2 882-3
[6] Gimpilevich Yu B, Vertegel V V and Tischuk Yu N 2020 Mathematical model of the measuring procedure of an automatic device for built-in monitoring of microwave path parameters Proc. of the 7th All-Russian Microwave Conf. (Moscow) 304-6
[7] Gimpilevich Yu B and Tischuk Yu N 2020 Calibration procedure for a quadrature meter of microwave path parameters Proc. of the 7th All-Russian Microwave Conf. (Moscow) 313-5
[8] Gimpilevich Yu B, Tischuk Yu N and Afonin I L 2020 Development, modeling and experimental research of a two-probe microstrip impedance sensor Proc. of the 7th All-Russian Microwave Conf. (Moscow) 310-2
[9] Gimpilevich Yu B and Zebek S E 2020 Proc. of the 7th All-Russian Microwave Conf. (Moscow) 307-9