2.45 GHz Band Quadrature Microwave Frequency Discriminators with Integrated Correlators Based on Power Dividers and Rat-Race Hybrids

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Abstract: Instantaneous frequency measurement devices are useful for conducting extremely fast measurements of the current frequency value of microwave signals, even if their duration is extremely short. This paper presents the principle of determination of temporary values of the microwave signal phase and frequency using interferometer techniques, based on passive microwave components. Additionally, the structures and results of measurements of two novel versions of integrated microwave correlators for microwave frequency discriminators, made on a single printed circuit board, are shown. Three Wilkinson-type, single-stage power dividers, and two rat-race hybrids create the developed correlators. The developed devices were designed to work over a wide frequency range, i.e., of 1.6–3.1 GHz, and can be used to monitor Wi-Fi devices as well as pulse and CW radar systems operating in the S band. They can also be applied in passive radars and active Doppler radars. The view of the printed circuits boards and results of measurements are presented. Recommendations for improving the accuracy of measurement are proposed.

Keywords: microwave frequency discriminator; microwave phase discriminator; microwave correlator; rat-race hybrid; electronic warfare; passive radar

1. Introduction

Microwave frequency discriminators (MFD) and microwave phase discriminators (MPhD) are the fundamental elements of instantaneous frequency measurement (IFM) systems [1–10]. Moreover, MFDs and MPhDs are attractive for phase noise measurements of microwave oscillators [11–16]. They can also work as simple demodulators in microwave FM receivers [17,18] and in monopulse direction finding systems of microwave emitters [19] and can be used for frequency identification [20]. Small-size microwave frequency discriminators can also work as blocks of passive radars for fast measurement of temporary values of parameters (pulse duration, temporary frequency, phase, and strength) of signals emitted by so-called emitters of opportunity. The MFD operating principle is based on a measurement of phase difference of microwave signals, propagating through two different lengths of transmission lines. Such frequency-sensitive devices belong to the group of the delay-line discriminators [7–10,12,21–25]. This phase difference is performed by the so-called proportional phase shift forming network (PPhSFN). The measurement of the phase difference is performed by the microwave phase discriminator, which is fed by the PPhSFN. The main segment of MPhD is a microwave multiport (six-port, for instance), which consists of microwave splitters and combiners [7–9,14,18,19,26]. There are also known quite simply as so-called single-function microwave frequency discriminators (SFMFD). Such devices are built of one microwave power splitter (power divider or directional coupler) and one microwave vector combiner or mixer [1,15,16,23–25,27–29] or use several band-stop microwave filters [30]. These types of instruments, i.e., SFMFD, can be used in cases when the accuracy of a frequency measurement is not a main requirement,
but their small size plays a particularly significant role. High measurement resolution and accuracy are the parameters that characterize the so-called quadrature microwave frequency discriminators (QMFD) [1–5,7–10,14]. The QMFDs may contain, depending on the implementation, several microwave splitters and at least two vector summation devices or mixers. In [1], for example, there was one model of QMFD described, which was based on only five rat-race couplers and there was also one version of QMFD comprised of five 90° overlap couplers presented. The QMFD with four 180° hybrids and one 90° (quadrature) hybrid was presented in [31]. The 90° hybrid in this discriminator was of the branch coupler type. The 180° hybrids were of the reverse-phase ring type. The same constructions of the 180° hybrids were used in the microwave phase discriminator described in [32]. In [1,5], a quadrature microwave frequency discriminator consisting of two power dividers and three 3 dB/90° directional couplers was presented. The quadrature frequency discriminator shown in [4] consists of three 3 dB/90° directional couplers, one 3 dB/180° directional coupler and one power divider. Research is currently underway to improve constructions of MFD. Modern discriminators should work as accurately as possible, be small in size, and not require complex technologies for their manufacturing. The unique pattern of integrated quadrature microwave frequency discriminator (i.e., proportional phase shift forming network and six-port on a single PCB) was developed by authors, and shortly presented in conference paper [33]. This QMFD was designed to run over one of the Wi-Fi frequency bands. In [33], the authors showed the structure of the integrated QMFD correlator and additionally presented measured transmission characteristics and reflection coefficients of all ports of this device.

Works on MFDs and MPhDs and the six-port devices for them is still underway, because they are very much needed, among others, in measurements and radiolocation, especially for monitoring the temporary values of the parameters of short-term microwave signals and signals with intense internal modulations. MFDs can measure not only the instantaneous frequency of the microwave signal, but also its instantaneous power. This measurement is available without the well-known down-conversion heterodyne techniques, even in very high (for example, millimeter-waves) frequency ranges [18,26]. Metallic guide and planar microstrip technologies are used for the MFDs and MPhDs fabrication [26]. Microwave frequency discriminators capable of measuring the instantaneous frequency value can be used in place of surface acoustic wave (SAW) resonators for wireless resonant sensors [7]. Ref. [8] shows the use of MFD in a frequency synthesizer as a stabilizing circuit instead of a phase locked loop (PLL). MFDs and MPhDs as parts of reconnaissance receivers can be used to create the so-called pulse description words (PDW) containing the parameters of the friend and foe microwave emitters [10].

The aim of the work is to develop microwave frequency discriminator with integrated correlator without crossings of internal connecting lines, operating in the S band. The production technology of Wilkinson power dividers and rat-race hybrids is simple and cheap. This caused the authors to look for solutions for the correlators under development using these types of components. Broadband phase and frequency discriminators were of the greatest importance for the works that were carried out.

Apart from the integrated correlator presented previously by the authors, the modification of this device, executed in the way of elimination of one of the components, is demonstrated in the paper. In addition, a more detailed mathematical description and principle of operation of the correlator for the quadrature microwave frequency discriminator is presented. This knowledge will enable readers to better understand physical phenomena and facilitate the design process of both integrated versions of microwave frequency discriminators and microwave phase discriminators. The measured input frequency–measured frequency transfer characteristics and measurement errors of microwave frequency discriminators with two developed versions of correlators were shown for comparison. The QMFDs, based on the developed integrated correlators, consisting of three power dividers and two rat-race hybrids, can largely meet requirements regarding the width of operation frequency band and measurement accuracy. A frequency band-
width of the developed as well as fabricated correlators for QMFD is close to one octave, even though that the operating width of frequency band of rat-race couplers is less than one octave.

The main novelty of the work is, in the authors’ opinions, the original structure of the microwave correlator for microwave frequency discriminator. Similar devices are known but not the same as that which we propose. The advisable structures of the correlators give designers and researchers further opportunities to build effective devices for instantaneous frequency and phase measurement systems.

2. Principle of Operation of Interferometer for a Quadrature Microwave Frequency Discriminator

The principle of operation of the interferometer for a quadrature microwave frequency discriminator is based on the vector summation of two signals $u_1$ and $u_2$ with the same amplitudes, $A$, the same frequencies, $f$, and with phases differing by the value of $\Theta$ proportional to the frequency, $f$. Signals $u_1$ and $u_2$ can be expressed with Equations (1) and (2), as follows:

$$u_1 = A \sin(2\pi f t + \Theta)$$

$$u_2 = A \sin(2\pi ft)$$

This phase difference $\Theta$ is created by the so-called proportional phase shift forming network, which is built of a power splitter (two-way power divider, for instance) and two transmission lines DL and TL with different physical lengths. The mathematical relation between signal frequency ($f$), phase difference ($\Theta$), light velocity ($c$), length of the DL line, marked as $L_{DL}$ and length of the TL line, marked as $L_{TL}$ (since they are TEM lines described with dielectric constant of $\varepsilon$) can be expressed according to a well-known Equation (3), as follows:

$$\Theta = 2\pi f c^{-1} (L_{DL} - L_{TL}) \varepsilon_{r}^{1/2}$$

As a vector summation component of microwave signals, a 3 dB quadrature directional coupler, such as a branch-line hybrid or Lange-type coupler, can be used. These elements, being four-port elements, execute two slightly different vector summation operations in parallel, in which the one of summation components (signals) is additionally shifted in phase by $90^\circ$. The symbol of a 3 dB quadrature directional coupler (DC) working as a vector summation device is shown in Figure 1.

![Figure 1](image)

Figure 1. A 3 dB directional coupler working as a vector summation component of microwave signals $u_1$ and $u_2$, (a) without additional phase shifting and (b) with additional phase shifting of signal $u_2$.

As a result, based on the two input signals $u_1$ and $u_2$ with a phase difference ($\Theta$) dependent on the unknown frequency ($f$), two microwave signals $u_a$ and $u_b$ are obtained with amplitude values of $A_a$ and $A_b$, respectively (Figure 1a). The values of $A_a$ and $A_b$ may not be the same. They are determined by the searched frequency ($f$) of the input microwave signals and can be described by Equations (4) and (5), as follows:

$$A_a = \left[ A^2 - A^2 \sin(\Theta) \right]^{1/2}$$

$$A_b = \left[ A^2 - A^2 \cos(\Theta) \right]^{1/2}$$
\[ A_b = \left[ A^2 + A^2 \sin(\Theta) \right]^{1/2} \]  

Based on Equations (4) and (5), it is possible to estimate the phase difference (\( \Theta \)) with Equation (6), as follows:

\[ \Theta = \arcsin \left( \frac{A_b^2 - A_a^2}{A_b^2 + A_a^2} \right) \]  

According to the properties of the arcsine function, Equation (6) facilitates the unambiguous determination of the change of the phase difference (\( \Theta \)) in the range of \( \pm \pi/2 \) radians. This situation limits the resolution of the frequency measurement. Therefore, before summing signals \( u_1 \) and \( u_2 \), by means of directional coupler, one of these signals should be shifted in phase by an additional 90° using phase shifter PhS (Figure 1b). In this way, the amplitudes \( A_c \) and \( A_d \) of output signals \( u_c \) and \( u_d \), resulting from the vector summation, can be described by Equations (7) and (8), as follows:

\[ A_c = \left[ A^2 - A^2 \cos(\Theta) \right]^{1/2} \]

\[ A_d = \left[ A^2 + A^2 \cos(\Theta) \right]^{1/2} \]

Based on Equations (7) and (8), it is possible to work out Equation (9), which enables the estimation of the phase difference (\( \Theta \)), as follows:

\[ \Theta = \arccos \left( \frac{A_d^2 - A_c^2}{A_d^2 + A_c^2} \right) \]

Therefore, based on the above-mentioned vector summation, operations providing two pairs of amplitudes values, i.e., \( A_a \) and \( A_b \), as well as \( A_c \) and \( A_d \), it will be possible to determine the phase difference (\( \Theta \)) in the range of full angle, i.e., \( \pm \pi \) radians, by using Equation (10), as follows:

\[ \Theta = \arctan \left( \frac{A_b^2 - A_a^2}{A_d^2 - A_c^2} \right) \]

Hence, following Equations (3) and (10), it is possible to find the unknown frequency (\( f \)) of the microwave signal by using Equation (11), as follows:

\[ f = \frac{\Theta_c}{2\pi(L_{DL} - L_{TL})c_1^{1/2}} \]

Of course, the evaluation of the frequency (\( f \)), using Equation (11), should be preceded by the estimation of the phase difference (\( \Theta \)).

3. Methodology

The main subassemblies of the correlator of a delay-line quadrature microwave frequency discriminator (QMFD) are the proportional phase shift forming network (PPhSFN) and the correlator of the quadrature microwave phase discriminator (QMPhD). The general structure of QMFD is shown in Figure 2. PPhSFN consists of a power divider (PD) and transmission lines DL and TL of unequal length. The PPhSFN provides a phase difference (\( \Theta \)) which is proportional to the frequency of the microwave signal reaching input port RF. The QMPhD correlator is a six-port microwave circuit, consisting of power splitters and vector summation components. The correlator of QMFD provides four microwave signals at its output ports 2, 3, 4, and 5. The power values \( P_1 \), \( P_2 \), \( P_3 \), and \( P_4 \) of these signals make
it possible to estimate the frequency of the input signal of the correlator. The frequency value is estimated according to Equation (12) and is denoted as \( f_m \).

\[
f_m = f_0 + \frac{c}{2\pi(L_{DL} - L_{TL})\epsilon_{ref}^{1/2}} \arctan \left( \frac{P_1 - P_2}{P_3 - P_4} \right)
\] (12)

**Figure 2.** Block diagram of the correlator for quadrature microwave frequency discriminator (general structure).

The value of \( f_0 \) is the center frequency of the measurement band of a correlator, i.e., when \( P_1 \) and \( P_2 \) are equal to each other. The so-called effective dielectric constant \( \epsilon_{ref} \) is used in Equation (12) when the correlator is fabricated in the microstrip technology. In Equation (12), the power values of the signals \( P_1, P_2, P_3, \) and \( P_4 \) were used instead of amplitudes, because microwave signal power can be measured more easily than the amplitude.

Equation (12) is applied when all QMFD correlator components are perfect, that is, perfectly matched, lossless, and have infinitely large isolations. When the components of the correlator are not perfect, it is necessary to perform a calibration of the QMFD to obtain the true input frequency–measured frequency transfer characteristics.

As power splitting elements needed in the design of correlator, Wilkinson-type power dividers are very suitable. As the vector summation components, the very good properties have Lange directional couplers. These couplers are quadrature in the phase domain and broadband and are relatively easy to implement in the lower bands of the microwave range. However, it is difficult to obtain their good parameters at higher frequencies. Another solution is using branch-line hybrids. Couplers of this type are also quadrature but the shapes of their amplitude characteristics over a wide frequency range differ from those of Lange couplers. Therefore, an alternative solution is to make a correlator for a quadrature microwave frequency discriminator, based on rat-race hybrids. The shapes of amplitude characteristic of the rat-race couplers in the middle of the operating band are like those of Lange couplers, although their bandwidth is narrower. Moreover, their phase parameters differ from the phase parameters of Lange couplers.

The rat-race hybrids belong to the of 0°/180° type couplers. As a result, rat-race couplers also perform two vector summation operations in parallel, but in one of the sums the phases of the signals are kept unchanged; whereas, in the other of the sums, one of the summed signals is phase shifted by 180 degrees. Thus, in fact, a signal obtained in one of the rat-race hybrid output ports is a vector difference of the input signals of this hybrid. In the developed correlator, this drawback was corrected by using appropriate lengths of transmission lines, connecting the relevant subassemblies.

Because of the cost and available technology, it was assumed that the correlator would be made with the use of FR4 substrate. The process of designing the QMFD correlator began with finding such lengths of rat-race hybrids branches, at which the center of their operating frequency band coincides with the required center of the QMFD operating band, i.e., 2.45 GHz. Then, the characteristic impedances of these branches were determined, for which the coupling coefficients amount to approximately −2.8 dB (slightly over −3 dB). The lengths of DL and TL lines were chosen so that the difference of their electric lengths at the frequency of 2.45 GHz was about 540 degrees. In subsequent design steps, the structure
of the phase shifter, the lengths of internal lines of the six-port and the DL and TL lines were searched for such that one of the waveforms of the transmittance modules to the port 2, 3, 4, or 5 had the shape of the transfer characteristic corresponding to the band-pass filter, and the remaining three transmittances had the shapes of the characteristics of the bandstop filters with appropriately shifted frequencies of maximum rejection. The reflection characteristics were also an important criterion for assessing the correlator at subsequent design stages.

4. Structure of the Developed Correlator for a Quadrature Microwave Frequency Discriminator

The design of the developed correlator, named the A correlator, based on rat-race hybrids [33], is presented in Figure 3. The first functional block of the correlator is the proportional (frequency–proportional) phase shift, forming network build of power divider PD1 and transmission lines DL and TL. The power divider PD1 equally splits the input signal feeding port 1 into two paths with lines TL and DL. Line DL is longer than line TL and the difference in the physical length of these lines can be marked as a delay line. The delay line length determines the measurement resolution and the width of the unambiguous frequency measurement band.

![Figure 3. Block diagram of the correlator for quadrature microwave frequency discriminator, based on power dividers and rat-race hybrids (correlator version A).](image-url)

The term unambiguous frequency measurement band, in this work, means the frequency range of the microwave input signal in which the transfer (discriminatory) characteristic of the quadrature microwave frequency discriminator is only increasing or only decreasing. The greater the difference in length of these lines, the better the frequency measurement resolution, but at the same time, the width of the unambiguous frequency measurement band is narrower. In turn, the power dividers PD2 and PD3, transmission lines L1, L2, Lref, L3, and L4, phase shifter PhS90°, and rat-race hybrids RRH1 and RRH2 create a six-port circuit, which is a correlator of microwave phase discriminator (MPhD). To obtain a sine dependence on the frequency of the signals power in ports 2 and 3, and a cosine dependence of the signals power in ports 4 and 5—as can be seen in Equations (4), (5), (7) and (8)—in the L4 line channel, the broadband PhS90° phase shifter was additionally placed to change the signal phase by a value of 90°. The parameter of the PhS90° phase shifter used in the developed correlator for frequency discriminator is the so-called differential phase shift [34–37]. The differential phase shift is the difference between the arguments of transmittances of the phase shifter and the adequate transmission line, which is called the reference line (marked as Lref in the block diagram shown in Figure 3). The Lref line is connected in cascade with the L2 transmission line. The power divider PD2 splits the signal received from the DL line output port into two signals, sent to hybrids RRH1 and RRH2, respectively. The power divider PD3 splits the signal from the TL line output port into two signals as well. One of them is sent to the hybrid RRH1 and the other to the phase shifter...
PhS90°. The hybrid RRH1 vectorially sums the signals from the PD1 and PD2 output ports. The hybrid RRH2 also sums up these signals, but the signal from line TL is additionally phase shifted by 90° using the phase shifter PhS90°. As a result, at the output ports of RRH1 and RRH2, two pairs of microwave signals are obtained. Their amplitudes (strengths) are the functions of the frequency and of parameters of the correlator components. Since the phase difference (Θ) resulting from the delay line is a function of the frequency (f) of the input signal (due to the difference in physical lengths of DL and TL lines), the signals strengths at the output ports 2, 3, 4, and 5 of the correlator are determined by the frequency of the signal fed to port number 1. The general relation between frequency (f), phase difference (Θ), light velocity (c), length of the DL line, and length of the TL line is given in Sections 2 and 3.

The correlator version A was designed, fabricated, and tested. The device was manufactured in microstrip technology using FR4 laminate with a thickness of 1.57 mm and dielectric constant of 4.6. The top view of the integrated correlator printed circuit board is shown in Figure 4. Its dimensions are 200 mm × 200 mm.

![Figure 4. Integrated correlator with two rat-race hybrids for quadrature microwave frequency discriminator (correlator version A).](image)

The rat-race hybrids RRH1 and RRH2 have ring shapes, and the power dividers PD1–PD3 are the Wilkinson single-stage devices. The phase shifter PhS90° has the form of single parallel open-circuited stub [34] with a length of 31.9 mm and width of 2 mm. The characteristic impedances of input and output lines, DL and TL lines, L1, L2, L3 L4, and Lref lines, are 50 Ohm. The widths of the paths of these lines are 2.73 mm. The widths of the paths of ring lines of the rat-race hybrids and power divider arms are 1.55 mm. The length of line DL is 180 mm, and the length of line TL is 60 mm, so the difference of their lengths is 120 mm. The lengths of lines L1–L4 are equal and are 60 mm. The length of the Lref line is 17 mm. All SMA connectors were mounted at the bottom side of the correlator’s PCB. Thus, the crossings of internal connecting lines were avoided.

In the case that the correlator shown in Figure 3 is used in the QMFD structure, then the tested signal with unknown frequency (f) is fed to port 1. Microwave power meters with analogue to digital (A/D) converters are connected to ports 2–5. The power values from them are used by the calculation unit to estimate the phase difference (Θ), and then for evaluation of the frequency of the microwave input signal feeding port 1. The evaluation algorithm of input signal frequency is based on (12) and employs discrimination characteristic of the correlator obtained in the calibration process.
5. Results of Measurements of Correlator Version A

The parameters of the developed correlator version A were measured using a vector network analyzer (VNA), model HP8720C HP. All of the 25 elements of the scatter matrix (S) of the device were measured. The results of measurements are presented in Figures 5–7.

![Figure 5](image1.png)

**Figure 5.** Reflection coefficients of ports of integrated correlator version A.

![Figure 6](image2.png)

**Figure 6.** Transmittances to non-isolated ports of the integrated correlator version A versus frequency.

![Figure 7](image3.png)

**Figure 7.** Transmittances between isolated ports of integrated correlator version A versus frequency.

The correlator A is characterized by reflection coefficients ($S_{ii}$) values (Figure 5) of each port less than $-10$ dB in the frequency range from about 2.1 GHz to about 3.1 GHz; whereas, $S_{ii}$ values better than $-7$ dB are in the range from 1.58 GHz to around 3.19 GHz (relative frequency bandwidth is slightly larger than one octave).

Maximum values of transmittances $S_{mn}$ to non-isolated ports—i.e., from port 1 to ports 2, 3, 4, and 5, which are presented in Figure 6—achieve about $-5$ dB. As is shown on the plots between vertical dashed lines in Figure 6 (i.e., in the frequency range from about 1.6 GHz to about 3.1 GHz), it is possible to unambiguously estimate the frequency of microwave signal feeding port 1.
Plots of modules of transmittances $S_{21}$, $S_{31}$, $S_{41}$, and $S_{51}$ have the shapes of characteristics of microwave band-pass filters set (bank) with center frequencies of about 2.75 GHz (in port 2), 2.05 GHz (in port 3), 2.45 GHz (in port 4), and 1.8 GHz, and 3.05 GHz (in port 5), respectively. Ports 2–5 can be also viewed as the outputs of band-stop filters. The channel with output port number 5, for example, works as a band-stop filter with center frequency of 2.45 GHz. The center frequencies of the other such band-stop filters are output port 2—about 2.09 GHz; output port 3—about 2.83 GHz; output port 4—about 1.72 GHz and about 3.19 GHz. These properties mean that such a correlator can also perform the function of a microwave four-channel multiplexer (single input and four outputs).

Ports 2, 3, 4, and 5 are mutually well isolated in the microwave frequency range. The transmittances between them are shown in Figure 7.

As it can be noticed, the isolations of ports: 4–2, 5–2, 4–3, and 5–3 are better than 12 dB over the entire tested frequency range. The traces of transmittances $S_{32}$ and $S_{54}$ have the shapes typical for 3 dB rat-race hybrids.

### 6. A Modified Structure of the Correlator with Rat-Race Hybrids for the Quadrature Microwave Frequency Discriminator (Correlator Version B)

One of the ways to decrease the size of any device is to limit the number of components used. In the middle of the operating frequency band of the developed correlator version A, the differential phase shift of the PhS90° shifter, considering the Lref line, is +90 degrees. In turn, the phase shift (argument of transmittance); added by reference line Lref; at the center frequency is −90 degrees. Therefore, it has been found that a +90 degree phase shift in the path of line L4 is equivalent to a −90 degree phase shift in path of line L2. Such a conclusion would indicate the possibility of eliminating the parallel stub forming the PhS90° shifter, while keeping the Lref line cascaded with the L2 line in the correlator. The simulations confirmed this hypothesis. Thus, in the second version of correlator (correlator B), the branch forming the PhS90° shifter was removed, but the Lref line was left as is. The block diagram of the correlator without the PhS90° phase shifter is shown in Figure 8. The view of the PCB of the modified correlator structure is shown in Figure 9.

![Figure 8](image-url). Scheme of the modified version of correlator (version B), based on rat-race hybrids for quadrature microwave frequency discriminator.
The principle of operation of this device does not change much in relation to the description presented above for correlator A. Similarly to the device shown in Figures 3 and 4, the power waveforms of microwave signals in ports 2 and 3 as well as 4 and 5 in the domain of frequency \( f \) are described with functions of the type of \( \sin(f) \) and \( \cos(f) \), respectively.

7. Results of Measurements of Integrated Correlator Version B

The results of experimental research are presented in Figures 10–12. They confirmed the correctness of the theoretical considerations mentioned above. The reflection coefficient (Figure 10) of the input of the modified correlator shown in Figure 9 increased from \(-10 \text{ dB}\) to about \(-8 \text{ dB}\), close to the frequency of 2 GHz. In turn, the reflection coefficient in port 5 improved so that in the entire tested frequency band did not exceed 12 dB.

![Figure 9](image-url)  
Figure 9. Modified version of integrated correlator (version B) with two rat-race hybrids for quadrature microwave frequency discriminator.

![Figure 10](image-url)  
Figure 10. Reflection coefficients of ports of the version B of integrated correlator versus frequency.
8. Results of Measurements of Microwave Frequency Discriminators with Integrated Correlators A and B

Microwave frequency discriminators, based on the developed correlators version A and B, were measured on the test bed shown in Figure 13. The measurement setup consists of correlator with four power meters connected to ports 2, 3, 4, and 5. The power meters are controlled by a computer with a program for estimating the temporary values of the frequency of the microwave signal fed to port 1.
are controlled by a computer with a program for estimating the temporary values of the correlator with four power meters connected to ports 2, 3, 4, and 5. The power meters and B, were measured on the test bed shown in Figure 13. The measurement setup consists of measurements (Figure 11). The constant coefficients used in Equations (13) and (14) were found in the calibration processes of the microwave frequency discriminators with the correlators that were developed. In these cases, it was assumed that the input frequency–measured frequency transfer characteristics of the microwave frequency discriminators, working with the correlators in the initial and modified versions (i.e., version A and B). In both cases, the measured frequency \( f_m \) (\( f_{mA} \) in case of correlator A and \( f_{mB} \) in case of correlator B) of the input signal in GHz was calculated basing on the power values \( P_1, P_2, P_3, \) and \( P_4 \) of the signals from the output ports 2–5 of both the correlators. For microwave frequency discriminator with correlator version A, the value of frequency \( f_m \) was calculated by Equation (13); whereas, for frequency discriminator with correlator version B, the value of frequency \( f_m \) was calculated by means of Equation (14).

\[
f_{mA} = 2.4622 + \frac{0.83467}{\pi} \cdot \arctan \left( \frac{P_1 - P_2}{P_3 - P_4} \right) \text{ [GHz]} \tag{13}
\]

\[
f_{mB} = 2.457 + \frac{0.68853}{\pi} \cdot \arctan \left( \frac{P_1 - P_2}{P_3 - P_4} \right) \text{ [GHz]} \tag{14}
\]

Figure 14 presents two examples of measured input frequency–measured frequency transfer characteristics of the microwave frequency discriminators, working with the correlators in the initial and modified versions (i.e., version A and B). In both cases, the measured frequency \( f_m \) (\( f_{mA} \) in case of correlator A and \( f_{mB} \) in case of correlator B) of the input signal in GHz was calculated basing on the power values \( P_1, P_2, P_3, \) and \( P_4 \) of the signals from the output ports 2–5 of both the correlators. For microwave frequency discriminator with correlator version A, the value of frequency \( f_m \) was calculated by Equation (13); whereas, for frequency discriminator with correlator version B, the value of frequency \( f_m \) was calculated by means of Equation (14).

\[
f_{ma} = 2.4622 + \frac{0.83467}{\pi} \cdot \arctan \left( \frac{P_1 - P_2}{P_3 - P_4} \right) \text{ [GHz]} \tag{13}
\]

\[
f_{mb} = 2.457 + \frac{0.68853}{\pi} \cdot \arctan \left( \frac{P_1 - P_2}{P_3 - P_4} \right) \text{ [GHz]} \tag{14}
\]

Figure 14. The input frequency–measured frequency transfer characteristics of quadrature microwave frequency discriminators, based on correlator version A and correlator version B (results of measurements).

The constant coefficients used in Equations (13) and (14) were found in the calibration processes of the microwave frequency discriminators with the correlators that were developed. In these cases, it was assumed that the input frequency–measured frequency transfer characteristics of the microwave frequency discriminators would be approximated by the single straight lines. As can be seen (Figure 14), the resulting transfer characteristics of QMFD with correlator A and correlator B are not the same but are remarkably similar.

Consequently, it can be concluded that the quality of the microwave frequency discriminator with the modified correlator is not worse than that with the developed correlator before modification, but it can be somewhat smaller.
The measurement error ($\Delta f_m$) of microwave frequency discriminators was calculated according to Equation (15), where $f_{inp}$ is the frequency value of the input signal of the microwave discriminator.

$$\Delta f_m = f_m - f_{inp}$$ (15)

The measurement errors of both microwave frequency discriminators were presented in Figure 15. For an input frequency value between 1.9 and 3.1 GHz, measurement error ($\Delta f_m$) was smaller than $\pm 25$ MHz for both developed discriminators. For $f_{inp} = 1.75$ GHz error ($\Delta f_m$) is about 56 MHz for discriminator with correlator A and error ($\Delta f_m$) is about 104 MHz for discriminator with correlator B.

![Figure 15. The measurement errors ($\Delta f_m$) versus input frequency ($f_{inp}$) of quadrature microwave frequency discriminators, based on correlator version A and correlator version B (results of measurements).](image)

Such accuracy is enough for many applications. When better measurement accuracy is needed, then in frequency estimation algorithms the authors use the frequency–measured frequency transfer characteristics, approximated by multiple (128 for instance) straight sections with appropriately selected slopes and constant factors.

9. Discussion

Integrated correlators for quadrature microwave frequency discriminators, based on two rat-race hybrids and three Wilkinson-type power dividers have been designed, fabricated, and successfully tested. Rat-race hybrids require slightly larger substrates than Lange couplers, for instance, but not very advanced technology is needed to fabricate them. In the case described here, the size of PCB of the modified integrated correlator was not reduced in order to be able to experimentally verify the correctness of the concept of dropping the parallel line branch, working as a phase shifter $\Phi S90^\circ$, without losing the properties of the correlator as a part of the quadrature microwave frequency discriminator. Even using the same FR4 substrate, the dimensions of this correlator (modified version B) can be reduced by about 40 percent in relation to the devices shown in Figures 4 and 9. Such devices are vital for investigations on rapid phenomena in microwave signals.

The use of thinner laminates and with a smaller dielectric constant will allow for a significant miniaturization of the developed correlators, if there is such an expectation.

As for future research directions, the authors anticipate works on limiting the deformation of QMFD transfer characteristics by reducing the size of the PCB and correcting the length and characteristic impedances of the individual transmission lines of the correlator and branches of the rat-race hybrids. A separate topic of the work will be the development of devices and algorithms for processing power values of output signals from correlators of microwave frequency discriminators, microwave phase discriminators for passive radars, and passive monopulse direction finding systems of dangerous microwave emitters.
10. Conclusions

The developed correlators, after a little correction, can work as the essential unit of quadrature microwave frequency discriminators in the frequency range from about 1.7 GHz up to about 3.4 GHz, so the relative frequency bandwidth is one octave.

These devices and their modifications can be used to measure and monitor radar systems and Wi-Fi devices operating in the S band.

The presented microwave frequency discriminators can work as an important block of instantaneous frequency measurement devices for electronic warfare applications and for measurements of internal frequency modulation of microwave signals emitted by CW radars, pulse radars, Doppler radars, and other telecommunications transmitters. The developed microwave discriminators can also work as signal identification blocks in the reference signal channels in passive radars that use CW or pulse signals. The six-port unit of the developed QMFD correlator, based on rat-race hybrids, can also work as a correlator of quadrature microwave phase discriminators (QMPhD) for small-size radars and for monopulse direction finding systems of microwave emitters. In such an application, the PhS90° phase shifter cannot be removed from the six-port structure shown in Figures 3 and 4 if a wide operating frequency range is required.

It has been proved that a single-channel correlator, made with the use of rat-race hybrids, can achieve a bandwidth at least of one octave.

The development and fabrication of the correlator of the MFD in an integrated form on one PCB allows for obtaining IFM devices of small sizes and in multi-channel versions. Each MFD channel will use a delay line of a different length.

MDFs and MPhDs are still needed. They are an alternative to microwave measurements made using conventional frequency conversion techniques [18,26].

The main conclusion is that studies on microwave frequency discriminators and microwave phase discriminators should be continued, as such networks have a wide range of applications. They can be used in measurements of microwave signals parameters, in vector network analyzers and, for example, in Doppler radar systems for detecting moving objects.

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