The Six-Port as a Communications Receiver

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***Abstract*—Configurable radio terminals require receivers with wide-band capabilities in order to support as many services as possible at most different carrier frequencies. Conventional well-known receiver architectures employing active circuitry are limited in this respect. Therefore, alternative architectures are investigated, such as the six-port, which has been introduced as a very flexible and elegant means for microwave measurements in the 1960s and 1970s. Later on, it has been used in radar applications. It was not until recently that communications receivers have been built upon the six-port principle. However, in all publications, there is always a certain mystic about the six-port. It has even been described as a "black box." In order to help paving the way for a wider application of the six-port technology, this paper describes the basic six-port theory and sets it into relation with the conventional receiver architectures such as the homodyne and heterodyne receiver. Finally, the advantages and possible applications of receivers based on the six-port technology are discussed.**

***Index Terms*—Communication terminals, frequency conversion, heterodyning, homodyne detection, receivers.**

I. Introduction

The requirements on wireless communications receivers are twofold. On one side, size and cost should be brought down; on the other side, the receivers should be more and more wide-band in order to meet the increasing demand for high data rates. Conventional heterodyne receivers require filters at the RF and IF, which can usually only be implemented by bulky surface acoustic wave (SAW) or crystal filters. Hence, an integration on a single chip and, thus, small size and low cost, cannot be achieved. For that reason, designers have been directing their effort toward the homodyne receiver (better known as the direct conversion receiver). The direct conversion receiver has gathered much attention, particularly within the research community since then. There are many problems due to analog impairments to be solved when implementing a direct conversion receiver. Therefore, other receiver architectures are sought after. One promising architecture is the six-port.

The application of the six-port to communications receivers has been presented in [1]–[6]. The general problem with these publications is the fact that the six-port itself is not explained. This is not satisfying. For many engineers who are only familiar with the above-mentioned conventional receiver architectures, it would be extremely useful to know the relationship between the six-port and conventional architectures. Instead of providing this knowledge, there is a certain mystic about the six-port. In [1] and [4], statements can be found such as "A six-port is a black box with two inputs and four outputs."

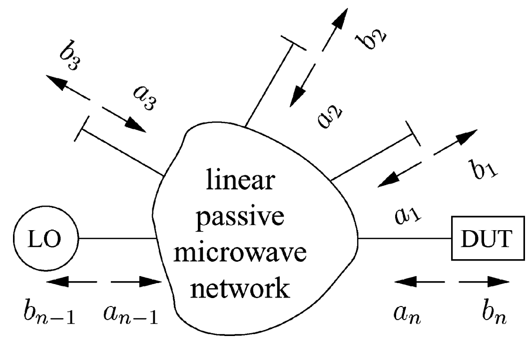

Fig. 1. Microwave network for the determination of the reflection coefficient of the DUT through independent remote measurements.

Among the first publications on the application of the six-port to communications receivers was [5]. Li *et al.* proved the concept of calculating the complex ratio of an incoming signal and a known local oscillator (LO) signal by using the six-port. Originally, the six-port was found to be a very good means to measure the complex reflection coefficient of a microwave device.

II. Background of the Six-Port Technique

The problem of microwave measurement is that connecting a probe to the device-under-test (DUT) considerably changes the very characteristics to be measured of the device, e.g., its reflection coefficient. This can be circumvented by determining the reflection coefficient of the DUT through a certain set of independent remote observations from a linear network to which the DUT is connected. Given the network of Fig. 1, it is possible to relate all "reflected" waves to all "incident" waves by means of the S-parameters

$$b_i = \sum_{j=1}^{n} S_{ij} a_j, \qquad i = 1, 2, \ldots, n \tag{1}$$

where S_{ij} are complex parameters and $a_i = |a_i| e^{j\varphi_{a_i}}$ and $b_i = |b_i| e^{j\varphi_{b_i}}$ are the complex amplitudes of the "incident" and "reflected" waves, respectively.

Assuming that the relationship of the "reflected" and "incident" waves at some of the ports (say, ports 1-m) can be uniquely described by the respective reflection coefficient Γ_i, it is

$$a_j = \Gamma_j b_j, \qquad j = 1, 2, \ldots, m \text{ and } m < n - 1. \tag{2}$$

Equations (1) and (2) form a system of $n+m$ equations with $2n$ unknown variables a_i and b_i. Hence, $n+m$ unknown variables can be described by the remaining $n-m$ variables. In the case

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of $m = n-2$, and by choosing a_n and b_n as the two unknown variables, it can be written in particular for b_i with $1 \le i \le m$

$$\begin{bmatrix} b_1 \\ b_2 \\ \vdots \\ b_{n-2} \end{bmatrix} = \begin{bmatrix} A_1 & B_1 \\ A_2 & B_2 \\ \vdots & \vdots \\ A_{n-2} & B_{n-2} \end{bmatrix} \begin{bmatrix} a_n \\ b_n \end{bmatrix} \tag{3}$$

where $A_i = |A_i|e^{j\varphi_{A_i}}$ and $B_i = |B_i|e^{j\varphi_{B_i}}$ are complex functions of the S_{ij} and Γ_i. Obviously, two observations b_1 and b_2 are sufficient to solve the system of equations for a_n and b_n given the proper choice of A_i and B_i. However, in practice, it is not a simple matter to measure the complex quantities b_i. Therefore, the idea arose to simply measure the power of the "reflected" waves b_i. Expanding the first row of (3) and calculating the power of b_1, it is

$$\begin{aligned} P_1 &= \frac{1}{2}|b_1|^2 \\ &= \frac{1}{2}|A_1 a_n + B_1 b_n|^2 \qquad (4) \\ &= \frac{1}{2}(A_1 a_n + B_1 b_n)(A_1 a_n + B_1 b_n)^* \\ &= \frac{1}{2}\Big(|A_1|^2|a_n|^2 + |B_1|^2|b_n|^2 + A_1 B_1^* a_n b_n^* \\ &\qquad + A_1^* B_1 a_n^* b_n\Big). \qquad (5) \end{aligned}$$

Equation (5) is linear in $|a_n|^2$, $|b_n|^2$, $a_n b_n^*$, and $a_n^* b_n$ [7]. Hence, four power observations are necessary to determine these quantities. The four observations and two additional ports (the DUT and LO) are the reason for the name "six-port." One of the first and also simplest six-ports presented in [8] for power measurements uses

$$\begin{bmatrix} A_1 & B_1 \\ A_2 & B_2 \\ A_3 & B_3 \\ A_4 & B_4 \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \\ 1 & 1 \\ 1 & j \end{bmatrix}.$$

The question might arise of whether a system of four equations of the type of (5) can always be solved. Clearly, the determinant of the coefficients must not be zero, which implies that at least one of the terms $A_i B_i^*$ is not real for $i \in \{1,2,3,4\}$. Consequently, the parameters A_i and B_i must not all be real or all imaginary. The above given example parameters fulfill this requirement.

Determining the quantities $|a_n|^2$, $|b_n|^2$, $a_n b_n^*$, and $a_n^* b_n$ does not provide a means to determine a_n and b_n. It is clear that $a_n b_n^*$ is the complex conjugate of $a_n^* b_n$. Both quantities comprise the same two parameters $|a_n||b_n|$ and $\arg(a_n) - \arg(b_n)$. Hence, with the help of $|a_n|^2$ and $|b_n|^2$, it is possible to determine $|a_n|$, $|b_n|$, and $\arg(a_n) - \arg(b_n)$, which can be used to determine the net power consumed in the DUT and the complex reflection coefficient of the DUT. Determining the absolute phase of a_n and b_n is not possible.

Since there are only three unknown parameters to be determined from (5), it is, in principle, sufficient to use a five-port with three power observations. The four quantities $|a_n|^2$, $|b_n|^2$, $a_n b_n^*$, and $a_n^* b_n$ of (5) are related to each other by $|a_n|^2 \cdot |b_n|^2 = a_n b_n^* \cdot a_n^* b_n$. Hence, reducing the solution to three equations leads to a system of nonlinear equations. Rewriting (5) yields

$$\begin{aligned} P_1 = \frac{1}{2}\Big(&|A_1|^2|a_n|^2 + |B_1|^2|b_n|^2 + 2|A_1||B_1||a_n||b_n| \\ &\times \cos\left(\varphi_{A_1} - \varphi_{B_1} + \varphi_{a_n} - \varphi_{b_n}\right)\Big). \end{aligned} \tag{6}$$

Assuming $|a_n|^2$ and $|b_n|^2$ can be obtained by two power observations P_2 and P_3 with $B_2 = 0$ and $A_3 = 0$, $\cos(\varphi_{A_1} - \varphi_{B_1} + \varphi_{a_n} - \varphi_{b_n})$ can be obtained by the observation P_1. Due to the nature of the cosine function, there are two solutions for $\varphi_{a_n} - \varphi_{b_n}$. Thus, the five-port does not yield a unique solution. The sixth port, i.e., the fourth power observation, can be used to choose among the two solutions. A geometric interpretation on this can be found in [9]. In [7], it is mentioned that the fourth power observation provides a means to improve and assess the measurement accuracy and, moreover, to simplify calibration procedures (i.e., the determination of the unknown parameters A_i and B_i).

It is clear that an equation of the type of (3) can be specified for any two unknown variables a_i and/or b_j. Of particular interest is a relationship between b_i ($i = 1, 2, \ldots, n-2$) on one side and a_{n-1} and a_n on the other side. This relationship describes the dependency between the "reflected" waves at the observation points and two "incident" waves. Further inspection reveals that all considerations made above with respect to a_n and b_n still hold when selecting a_{n-1} and a_n as the two unknown variables. Therefore, the equations do not need to be rewritten. The two unknown waves a_n and b_n can stand for any two unknown waves a_i and/or b_j. They are not limited to be the "incident" wave and the "reflected" wave of the DUT. This is of particular importance for the understanding of (24) and (25).

At this point, the classical six-port theory should be left by asking for the instantaneous power, i.e., a power signal $p_1(t)$ whose mean is the above-mentioned power P_1. To do so, the following two complex waves are introduced with generally different frequencies f_a and f_b

$$a_i(t) = a_i e^{j2\pi f_a t} \tag{7}$$

$$b_i(t) = b_i e^{j2\pi f_b t}. \tag{8}$$

It should be noted that these time-varying waves can be used instead of the complex amplitudes without any change in the above equations. Hence, the instantaneous power of the "reflected" wave of the second port is given by (9), shown at the bottom of the following page.

If $f_a = f_b$, the mean of $p_1(t)$ is clearly P_1. In the case of $f_a \neq f_b$, the high-frequency components can be removed by means of low-pass filtering (which can be interpreted as computing the short-term mean). Hence, the following signal can be obtained:

$$\begin{aligned} y_1(t) &= \mathrm{LP}\left(p_1(t)\right) \\ &= \frac{1}{2}|A_1|^2|a_n|^2 + \frac{1}{2}|B_1|^2|b_n|^2 + |A_1||B_1||a_n||b_n| \\ &\quad \times \cos\left(2\pi(f_a - f_b)t + \varphi_{A_1} - \varphi_{B_1} + \varphi_{a_n} - \varphi_{b_n}\right) \end{aligned} \tag{10}$$

where LP is an operator reflecting the low-pass filtering to remove the high-frequency components at $2f_a$, $2f_b$, and $f_a + f_b$. This signal contains dc components, as well as a component

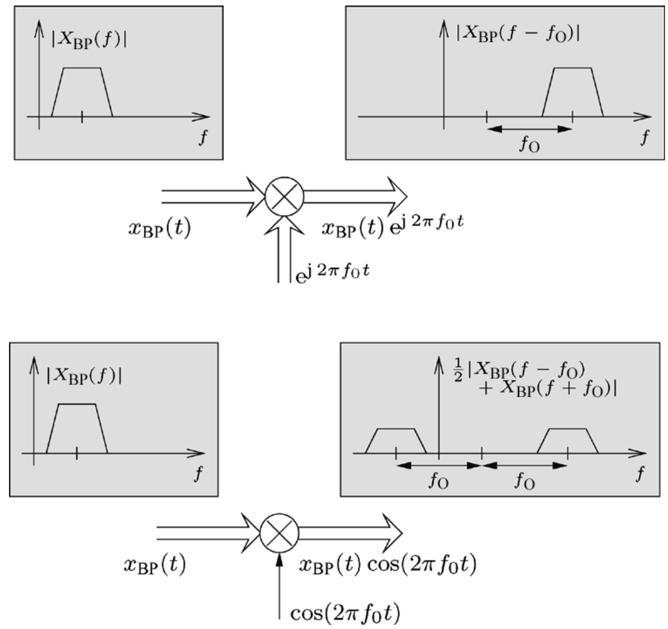

Fig. 2. (*top*) Complex and (*bottom*) real frequency-conversion principle.

at the frequency $f_a - f_b$. Obviously, a frequency-conversion process has taken place.

III. Frequency Conversion

Frequency conversion is the process of shifting the spectrum of a signal. There are several reasons for doing this, i.e., the separation of a certain number of signals on a transmission medium (using different carrier frequencies) or to be able to radiate a signal by an antenna.

Hence, the (Fourier) spectrum $X_{\rm BP}(f)$ of the (bandpass) signal $x_{\rm BP}(t)$ is shifted by f_0 resulting in $X_{\rm BP}(f-f_0)$. From the properties of the Fourier transform, it can easily be concluded that this frequency shift can be achieved by multiplying the signal $x_{\rm BP}(t)$ with $e^{j2\pi f_0 t}$ (see Fig. 2). One of the main problems with this theoretical approach is that the complex harmonic signal $e^{j2\pi f_0 t}$ must be realized in a LO by means of its real part and its imaginary part, i.e., $\cos(2\pi f_0 t)$ and $\sin(2\pi f_0 t)$. In practice, it is a problem to realize the exact phase difference of $\pi/2$ for this *complex frequency conversion*.

Therefore, and in order to reduce the effort, one could support the idea of *real frequency conversion* through multiplication of the signal $x_{\rm BP}(t)$ by $\cos(2\pi f_0 t)$ (or $\sin(2\pi f_0 t)$). Expanding the cosine function to $1/2(e^{j2\pi f_0 t} + e^{-j2\pi f_0 t})$ clearly reveals that the real frequency conversion results in two separate frequency shifts of the original spectrum $X_{\rm BP}(f)$ (see Fig. 2).

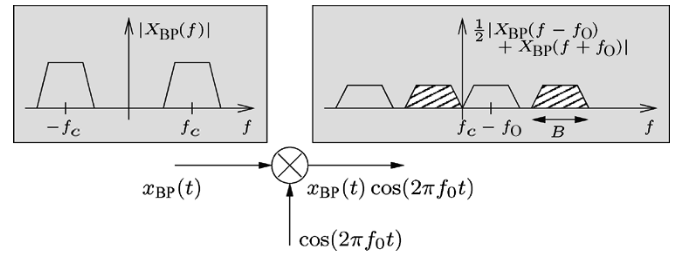

Fig. 3. Real frequency conversion of a real bandpass signal.

In case the signal $x_{\rm BP}(t)$ is a result of a real frequency-conversion process (e.g., up-conversion of a baseband signal to a carrier frequency in a transmitter), another real frequency conversion must be carried out with caution since the two sidebands of $x_{\rm BP}(t)$ might overlap after the second frequency conversion, as can be seen from Fig. 3. This can obviously be avoided if one of the following conditions is obeyed:

$$\frac{B}{2} < f_0 < f_c - \frac{B}{2}$$

or

$$f_0 > f_c + \frac{B}{2}. \tag{11}$$

In the special case, each of the two sidebands $x_{\rm BP}(t)$ is symmetric about f_c, a constructive overlap with $f_0 = f_c$ is, in principle, possible. However, care must be taken regarding the phase difference between the carrier $\cos(2\pi f_c t)$ and the LO. Multiplying $x_{\rm BP}(t) = x(t)\cos(2\pi f_c t)$ by $\cos(2\pi f_c t + \varphi)$ results in $1/2\, x(t)(\cos(\varphi) + \cos(2\pi 2 f_c t + \varphi))$. Depending on the value of φ, the baseband component $x(t)$ might be completely deleted. Hence, in this case, the real frequency conversion of the real bandpass signal must be phase synchronous (i.e., coherent).

In the case $x(t)$ is a complex signal, the two sidebands of $x_{\rm BP}(t)$ are not symmetric. Therefore, $x(t)$ can only be obtained from $x_{\rm BP}(t)$ through complex frequency conversion with $f_0 = f_c$. This can also be seen from very simple algebraic considerations. The real bandpass signal can be obtained from the complex baseband signal $x(t)$ through complex up-conversion of $x(t)$ to f_c, which yields a complex bandpass signal. Taking the real part results in a real signal $x_{\rm BP}(t)$ with a symmetric spectrum formed by two sidebands, which, taken separately, are not symmetric. Hence, it is

$$\begin{aligned} x_{\rm BP}(t) &= \mathfrak{Re}\left\{x(t)e^{j2\pi f_c t}\right\} \\ &= x_I(t)\cos(2\pi f_c t) - x_Q(t)\sin(2\pi f_c t) \end{aligned} \tag{12}$$

$$\begin{aligned} p_1(t) =& \left(\mathfrak{Re}\{b_1(t)\}\right)^2 \\ =& \left(|A_1||a_n|\cos(2\pi f_a t + \varphi_{A_1} + \varphi_{a_n}) + |B_1||b_n|\cos(2\pi f_b t + \varphi_{B_1} + \varphi_{b_n})\right)^2 \\ =& \frac{1}{2}|A_1|^2|a_n|^2(1+\cos(2\pi \boldsymbol{2 f_a} t + 2\varphi_{A_1} + 2\varphi_{a_n}) + \frac{1}{2}|B_1|^2|b_n|^2(1+\cos(2\pi \boldsymbol{2 f_b} t + 2\varphi_{B_1} + 2\varphi_{b_n}) \\ &+ |A_1||B_1||a_n||b_n|\left(\cos(2\pi(f_a - f_b)t + \varphi_{A_1} - \varphi_{B_1} + \varphi_{a_n} - \varphi_{b_n}) + \cos(2\pi(\boldsymbol{f_a} + \boldsymbol{f_b})t + \varphi_{A_1} + \varphi_{B_1} + \varphi_{a_n} + \varphi_{b_n})\right) \end{aligned} \tag{9}$$

where $x_I(t)$ is the real part of $x(t)$, and $x_Q(t)$ is the imaginary part of $x(t)$. These two components of $x(t)$ are also called the *in-phase* (I) and *quadrature-phase* (Q) components of $x(t)$.

Hence, in order to obtain the full information carried by the complex signal $x(t)$, one has to have the two real signals $x_I(t)$ and $x_Q(t)$. However, performing real down-conversion yields only one real signal. Real down-conversion and low-pass filtering yields

$$\begin{aligned} x_M(t) &= \mathrm{LP}\big(x_{\mathrm{BP}}(t)\cos(2\pi f_c t + \varphi_1)\big) \\ &= \mathrm{LP}\big(x_I(t)\cos(2\pi f_c t)\cos(2\pi f_c t + \varphi_1) \\ &\qquad - x_Q(t)\sin(2\pi f_c t)\cos(2\pi f_c t + \varphi_1)\big) \\ &= \frac{1}{2}\big(x_I(t)\cos(\varphi_1) + x_Q(t)\sin(\varphi_1)\big). \end{aligned} \tag{13}$$

The result is a mixture of $x_I(t)$ and $x_Q(t)$. Another "observation" $x_N(t)$ is necessary in order to be able to separate the two components. This can be obtained by a real down-conversion with $\cos(2\pi f_c t + \varphi_2)$. Consequently, it can be written

$$\begin{bmatrix} x_M(t) \\ x_N(t) \end{bmatrix} = \frac{1}{2}\begin{bmatrix} \cos(\varphi_1) & \sin(\varphi_1) \\ \cos(\varphi_2) & \sin(\varphi_2) \end{bmatrix}\begin{bmatrix} x_I(t) \\ x_Q(t) \end{bmatrix}. \tag{14}$$

The two components $x_I(t)$ and $x_Q(t)$ can be calculated from the two observations if the matrix is nonsingular. The (theoretically) classic case of complex down-conversion is $\varphi_1 = 0$ and $\varphi_2 = \pi/2$. However, it can be concluded from (14) that, as long as the phases are known and as long as the matrix of (14) can be inverted, the typically required phase difference of $\varphi_2 - \varphi_1 = \pi/2$ is, in principle, not necessary in order to obtain $x_I(t)$ and $x_Q(t)$.

IV. Homodyne and Heterodyne Principle

A. General

In a modern receiver, the incoming bandpass signal is usually down-converted to baseband. This can be realized in one or several steps. Depending on whether $f_0 = f_c$ or not, two types of receivers can be distinguished. Using $f_0 = f_c$ is intuitively the first choice if the task is to down-convert a bandpass signal to baseband. This approach is called the *homodyne* principle since the carrier frequency is equal (homo) to the frequency of the LO. In the case the two frequencies are different, it is called the *heterodyne* principle, which yields a signal at an IF. This signal is usually down-converted to baseband in a second frequency-conversion step. It is possible to use both principles in connection with real and complex frequency conversion.

B. Homodyne Receiver With Complex Frequency Conversion

Using (14) with $\varphi_2 - \varphi_1 = \pi/2$ yields, apart from a phase shift, the in-phase and quadrature-phase components of the signal $x(t)$. The corresponding receiver type is also called a *direct down-conversion receiver*. It is sketched in Fig. 4. In practice, the phase difference of $\varphi_2 - \varphi_1 = \pi/2 = 90^\circ$ cannot be realized exactly. Feasible values of the deviation from 90° lie between 1° and 3° [10], [11]. Moreover, there is usually a gain mismatch between the in-phase and quadrature-phase branches of a direct down-conversion receiver. The reader is referred to the chapter of Beach *et al.* in [12] for details on realization problems of direct down-conversion receivers.

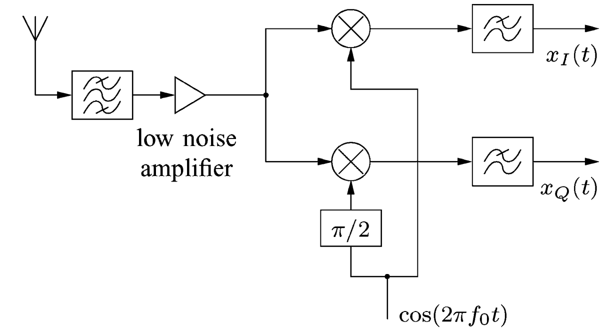

Fig. 4. Receiver with complex frequency conversion ($f_0 = f_c$: direct down-conversion receiver, $f_0 \neq f_c$: low-IF receiver).

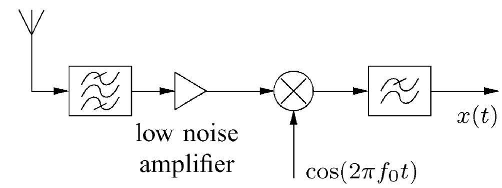

Fig. 5. Receiver with real frequency conversion (not showing the possibly required carrier recovery).

C. Homodyne Receiver With Real Frequency Conversion

From the simple algebraic considerations above, it is clear that a homodyne receiver with real frequency conversion is only feasible for real baseband signals. Real frequency conversion delivers one "observation." With (13), it can be concluded that either $x_I(t)$ or $x_Q(t)$ must be zero in order to obtain the transmitted signal from the received signal. Moreover, the coefficient $\cos(\varphi_1)$ or $\sin(\varphi_1)$, respectively, must not be zero. Therefore, homodyne receivers with real frequency conversion must be synchronous receivers. This has been discussed in Section III.

The synchronous LO can either be transmitted (e.g., the pilot reference in the conventional FM stereo broadcasting system) or be regenerated from the received signal (e.g., by a so-called Costas loop). A typical structure is shown in Fig. 5.

D. Heterodyne Receiver With Complex Frequency Conversion

Using the structure of Fig. 4 with $f_0 \neq f_c$ yields a heterodyne receiver with complex frequency conversion. In practice, f_0 is usually chosen very near to f_c, hence, the IF $f_{IF} = |f_c - f_0|$ is low. Consequently, these receivers are named *low IF receivers*. Their great advantage over the direct down-conversion receiver is that the dc offset (primarily caused by self-mixing due to finite isolation of the components) does not overlap with the wanted signal. Hence, it can be easily removed by means of filtering.

The disadvantage compared to the direct down-conversion receiver is the fact that the gain and phase mismatches have a more severe impact [12].

E. Heterodyne Receiver With Real Frequency Conversion

Using the structure of Fig. 5 with $f_0 \neq f_c$ yields a heterodyne receiver with real frequency conversion. Since there is only one branch of signal processing, no problems exist regarding gain or phase mismatches. However, as discussed earlier, it must be taken care that the individual frequency bands do not overlap after frequency conversion. This is guaranteed if (11) is obeyed. To ensure this, a bandpass filter might be required that limits the signal bandwidth to the smaller value of

$$B < 2f_0 \text{ and } B < 2|f_0 - f_c|. \tag{17}$$

Since the possibly overlapping frequency band of $x_{\mathrm{BP}}(t)$ (i.e., the shifted left-hand-side band of the real bandpass signal) is called the image (of the right-hand-side band), this filter is called the *image rejection filter*. Hence, the effort for the second mixer for complex frequency conversion is saved at the cost of an image rejection filter. Since this filter is usually an off-chip filter, e.g., a SAW filter, designers try to refrain from the real frequency conversion and try to make the complex frequency conversion feasible, thus enabling a higher degree of integration on a single chip.

V. Realization Aspects

Frequency conversion (mixing) is achieved by multiplying a signal with a real or a complex LO signal. Consequently, the straightforward realization is the application of multipliers or equivalent circuits that perform the multiplication. This type of the frequency conversion is called the *multiplicative mixing*.

Another means of realizing the frequency conversion is the *additive mixing*. The LO signal is added to the signal $x_{\mathrm{BP}}(t)$ and the resulting sum is nonlinearly processed (e.g., by using the current–voltage characteristics of a diode). The nonlinear characteristics in the operating point of a certain device can be approximated by a Taylor series with the constants $K_1, K_2, \ldots,$ as follows:

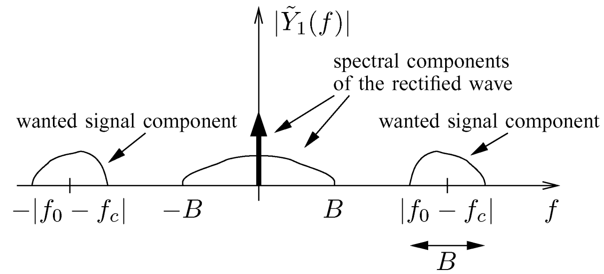

Fig. 6. Spectrum of a signal after additive mixing and low-pass filtering.

$$\tilde{p} = K_0 + K_1 q + \frac{K_2}{2} q^2 + \frac{K_3}{6} q^3 + \ldots.$$

If q is the superposition of $x_{\mathrm{BP}}(t)$ and the LO with an amplitude G_1 and a phase φ_1, it is

$$\tilde{p}_1(t) = K_0 + K_1\big(x_{\mathrm{BP}}(t) + G_1 \cos(2\pi f_0 t + \varphi_1)\big) + \frac{K_2}{2}\big(x_{\mathrm{BP}}(t) + G_1\cos(2\pi f_0 t + \varphi_1)\big)^2 + \ldots. \tag{18}$$

Using (12), (15), shown at the bottom of this page, can be derived. In the case of $K_i \approx 0$ for $i > 2$, the higher order harmonics and intermodulation products can be neglected. Assuming f_0 is relatively near to f_c, a signal according to (16), shown at the bottom of this page, can be obtained from low-pass filtering. The dc component K_0 of (15) has been neglected in favor of the rectified wave.

The spectrum of the signal $\tilde{y}_1(t)$ is sketched in Fig. 6. It should be noted that the spectral components of signals of the type $x_z^2(t)$ occupy twice the bandwidth of a low-pass signal $x_z(t)$ and comprise a single tone at dc.

$$\begin{aligned}
\tilde{p}_1(t) = {} & K_0 + K_1 \underbrace{\big(x_{\mathrm{BP}}(t) + G_1\cos(2\pi f_0 t + \varphi_1)\big)}_{\text{fundamental wave}} + \underbrace{\frac{K_2}{4}\left(x_I^2(t) + x_Q^2(t) + G_1^2\right)}_{\text{rectified wave}} \\
& + \underbrace{\frac{K_2}{2}\left(x_I(t) G_1 \cos\left(2\pi(f_0 - f_c)t + \varphi_1\right) + x_Q(t) G_1 \sin\left(2\pi(f_0 - f_c)t + \varphi_1\right)\right)}_{\text{difference frequency}} \\
& + \underbrace{\frac{K_2}{2}\left(x_I(t) G_1 \cos\left(2\pi(f_0 + f_c)t + \varphi_1\right) - x_Q(t) G_1 \sin\left(2\pi(f_0 + f_c)t + \varphi_1\right)\right)}_{\text{sum frequency}} \\
& + \underbrace{\frac{K_2}{4}\left(G_1^2\cos(2\pi 2 f_0 t + 2\varphi_1) + \left(x_I^2(t) - x_Q^2(t)\right)\cos(2\pi 2 f_c t) - 2x_I(t)x_Q(t)\sin(2\pi 2 f_c t)\right)}_{\text{first harmonic}} \\
& + \underbrace{\cdots}_{\text{higher order harmonics and intermodulation products}}
\end{aligned} \tag{15}$$

$$\begin{aligned}
\tilde{y}_1(t) &= \mathrm{LP}\left(\tilde{p}_1(t)\right) \\
&= \underbrace{\frac{K_2}{4}\left(x_I^2(t) + x_Q^2(t) + G_1^2\right)}_{\text{rectified wave}} + \underbrace{\frac{K_2}{2}\left(x_I(t) G_1 \cos\left(2\pi(f_0 - f_c)t + \varphi_1\right) + x_Q(t) G_1 \sin\left(2\pi(f_0 - f_c)t + \varphi_1\right)\right)}_{\text{wanted signal components}}
\end{aligned} \tag{16}$$

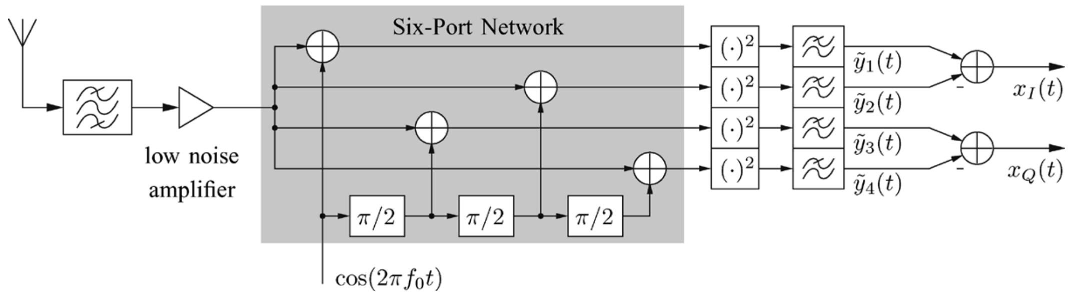

Fig. 7. Simple six-port receiver with analog I/Q generation.

The wanted components of $\tilde{y}_1(t)$ are the difference frequency components. In the case they do not overlap with the other components, they can be separated by means of filtering. This is possible if

$$f_0 < f_c - \frac{3}{2}B$$

or

$$f_0 > f_c + \frac{3}{2}B \tag{19}$$

which can be accomplished by carefully selecting f_0 or by band-limiting the signal before mixing with a filter having the bandwidth

$$B < \frac{2}{3}|f_0 - f_c|. \tag{20}$$

The basic structure realizing a heterodyne receiver with additive mixing shall be named the superposition heterodyne receiver or short *superheterodyne receiver*.[1] Setting $f_0 = f_c$ yields

$$\tilde{y}_1(t) = \frac{K_2}{4}\left(x_I^2(t) + x_Q^2(t) + G_1^2\right) + \frac{K_2}{2}\left(x_I(t)G_1\cos(\varphi_1) + x_Q(t)G_1\sin(\varphi_1)\right) \tag{21}$$

which very much resembles (13), except for the rectified wave component. By inspection, it can be found that subtracting a second observation $\tilde{y}_2(t)$ with $G_2 = G_1$ and $\varphi_2 = \varphi_1 + \pi$ from $\tilde{y}_1(t)$ yields something equivalent to $x_M(t)$ of (13). Consequently, by choosing $\varphi_1 = 0$ and another two observations with $\varphi_3 = \pi/2$, $\varphi_4 = \varphi_3 + \pi$, and $G_1 = G_2 = G_3 = G_4$, complex down-conversion to baseband can be achieved. The resulting receiver structure is sketched in Fig. 7 for $G_i = 1$. It uses four observations of a linear network with two inputs (the LO and signal $x_{\mathrm{BP}}(t)$). The four observations are processed by a square-law device (e.g., a diode). Hence, the structure of Fig. 7 is a six-port receiver with analog I/Q regeneration. It has been introduced in [1].

The analog I/Q regeneration relies on a perfect cancellation of the rectified wave component, which is only possible if the four branches are perfectly balanced and if the LO amplitudes G_i have the same magnitude. Since this is not realizable in practice, the receiver of Fig. 7 has certainly its limitations. Nonetheless, it proves the concept of a homodyne receiver employing additive mixing.

[1]The author is aware of the fact that the term *superheterodyne* originally stands for supersonic heterodyne due to the fact that the IF is supersonic. However, in times where the term supersonic is no longer meaningful in this context, it shall be allowed to have a new meaning.

VI. Five- and Six-Port Receivers

Critical inspection of Fig. 7 reveals that four observations are used, although (21) comprises only the three unknown $x_I^2(t) + x_Q^2(t)$, $x_I(t)$, and $x_Q(t)$ (assuming that K_2, G_1, and φ_1 are known).

Again, the algebraic interpretation of the problem is very helpful to generalize the problem. Using three power observations of the type of (21) and assuming that the constant factor K_2 is equal for all three branches yields

$$\begin{bmatrix}\tilde{y}_1(t)\\ \tilde{y}_2(t)\\ \tilde{y}_3(t)\end{bmatrix} = \frac{K_2}{4}\begin{bmatrix}2G_1\cos(\varphi_1) & 2G_1\sin(\varphi_1) & 1\\ 2G_2\cos(\varphi_2) & 2G_2\sin(\varphi_2) & 1\\ 2G_3\cos(\varphi_3) & 2G_3\sin(\varphi_3) & 1\end{bmatrix} \times \begin{bmatrix}x_I(t)\\ x_Q(t)\\ x_I^2(t)+x_Q^2(t)\end{bmatrix} + \frac{K_2}{4}\begin{bmatrix}G_1^2\\ G_2^2\\ G_3^2\end{bmatrix}. \tag{22}$$

The wanted signal components $x_I(t)$ and $x_Q(t)$ can be calculated if the matrix is nonsingular, which can be forced by properly selecting φ_i and G_i. A structure realizing (22) is shown in Fig. 8. It is a homodyne receiver employing additive mixing. Therefore, it shall be named the superposition homodyne receiver or short the *superhomodyne receiver*. The low-pass filters preceding the analog-to-digital (AD) converters serve as antialiasing filters, as well as to remove the high-frequency components of the signals and, thus, reflect the step from (15) to (16). The signal components $x_I(t)$ and $x_Q(t)$ are calculated in the digital domain by using the digitized observations $\tilde{y}_i$.

In the case that after I/Q regeneration another (digital) frequency conversion is applied (e.g., to select a certain subband of the signal x), the receiver is, strictly spoken, not a homodyne receiver. However, regarding x as the wanted signal that is converted to baseband, the term homodyne is justifiable.

To close the circle, the relationship between (10) and (16) is shown. To this end (16) is rewritten using the well-known relationships between $x(t)$, $x_I(t)$, and $x_Q(t)$ as follows:

$$x(t) = |x(t)|\,e^{j\varphi_x(t)} = x_I(t) + jx_Q(t)$$

which yields

$$\tilde{y}_1(t) = \frac{K_2}{4}\left(|x(t)|^2 + G_1^2\right) + \frac{K_2}{2}G_1\,|x(t)| \times\cos\left(2\pi(f_0 - f_c)t + \varphi_1 - \varphi_x(t)\right). \tag{23}$$

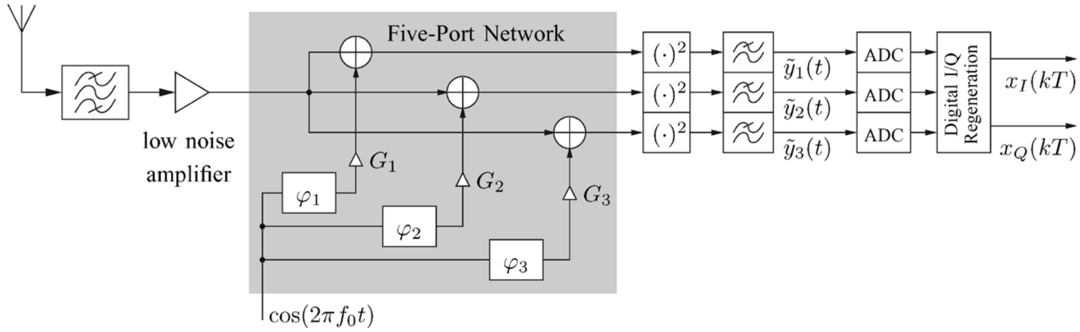

Fig. 8. Five-port receiver with AD converters and digital signal-processing block for regenerating the in-phase and quadrature-phase components of the received signal.

Neglecting the factor $K_2/2$ and comparing with (10) yields $\tilde{y}_i = y_i(t)$ with, for example, the following parameters:

$$\begin{matrix} a_n = 1 & A_i = G_i e^{j\varphi_i} \\ b_n = x(t) & B_i = 1 \end{matrix}, \qquad \text{for } i = 1, 2, 3. \tag{24}$$

Hence, the superhomodyne receiver is a special five-port (or six-port) receiver. Likewise, it is possible to get (10) with $f_a = f_b$ in the form of (22). Using the abbreviations $\Delta\phi_i = \varphi_{A_i} - \varphi_{B_i}$ and $\Delta\varphi = \varphi_{a_n} - \varphi_{b_n}$, this yields the arithmetic description of a somewhat more general superhomodyne receiver where b_n is assumed to be related to the received signal and a_n is assumed to be related to the LO (see Section II for the definition of a_n and b_n) as follows:

$$\begin{bmatrix} y_1(t) \\ y_2(t) \\ y_3(t) \end{bmatrix} = \begin{bmatrix} |A_1||B_1|\cos(\Delta\phi_1) & -|A_1||B_1|\sin(\Delta\phi_1) & \frac{1}{2}|B_1|^2 \\ |A_2||B_2|\cos(\Delta\phi_2) & -|A_2||B_2|\sin(\Delta\phi_2) & \frac{1}{2}|B_2|^2 \\ |A_3||B_3|\cos(\Delta\phi_3) & -|A_3||B_3|\sin(\Delta\phi_3) & \frac{1}{2}|B_3|^2 \end{bmatrix} \times \begin{bmatrix} |a_n||b_n|\cos(\Delta\varphi) \\ |a_n||b_n|\sin(\Delta\varphi) \\ |b_n|^2 \end{bmatrix} + \frac{1}{2}|a_n|^2 \begin{bmatrix} |A_1|^2 \\ |A_2|^2 \\ |A_3|^2 \end{bmatrix}. \tag{25}$$

Substituting (24) to (25) yields $\Delta\phi_i = \varphi_i$ and $\Delta\varphi = -\varphi_x(t)$ and, thus, establishes (22), except for the constant factor $K_2/2$.

It should be noted that, in practical applications, the right-hand-side term of (22) and (25), as well as components of $x_I^2(t) + x_Q^2(t)$ or $|b_n|^2$, respectively, are usually removed e.g., by means of high-pass filtering or by means of estimating the mean of the low-pass filtered instantaneous power observations and subtracting it.

It remains the question of how to obtain the wanted signal components $x_I(t)$ and $x_Q(t)$ from the three or four observations. A comprehensive treatment of this topic is beyond the scope of this paper. However, the following two basic approaches should be mentioned.

- The unknown parameters A_i and B_i are determined by means of a calibration procedure. Hence, given a proper choice of the parameters, the matrices of (22) and (25) can be inverted, which yields the desired signal components. A calibration technique for six-port receivers has been presented in [13].
- Instead of estimating the unknown parameters of the matrices, the unknown wanted signal components $x_I(t)$ and $x_Q(t)$ can be estimated. A very simple technique has been presented in [14]. In [15] and [16], a technique has been presented to separate the signal subspace spanned by $x_I(t)$ and $x_Q(t)$ from the noise subspace. The wanted signal components can be obtained from the signal subspace by conventional phase synchronization techniques. However, the research in [15] and [16] is based on assumptions that are not justifiable, as has been shown in [14].

However, it is not always necessary to perform a digital I/Q regeneration as the analog I/Q regeneration of Fig. 7 suggests. Similar structures have been presented in [4] and [17]. In [18], a five-port receiver with analog I/Q regeneration is presented. The idea behind it is to properly choose the parameters of (22), e.g., $G_1 = G_2 = G_3 = G$ and $\varphi_1 = \pi/4$, $\varphi_2 = 3\pi/4$, and $\varphi_3 = 7\pi/4$. The wanted signal components can be calculated as

$$\begin{bmatrix} \tilde{x}_I(t) \\ \tilde{x}_Q(t) \end{bmatrix} = \begin{bmatrix} 1 & -1 & 0 \\ 1 & 0 & -1 \end{bmatrix} \begin{bmatrix} y_1(t) \\ y_2(t) \\ y_3(t) \end{bmatrix}$$
$$= \frac{K_2}{4} \begin{bmatrix} 1 & -1 & 0 \\ 1 & 0 & -1 \end{bmatrix} \times \begin{bmatrix} 2G\cos\frac{1}{4}\pi & 2G\sin\frac{1}{4}\pi & 1 \\ 2G\cos\frac{3}{4}\pi & 2G\sin\frac{3}{4}\pi & 1 \\ 2G\cos\frac{7}{4}\pi & 2G\sin\frac{7}{4}\pi & 1 \end{bmatrix} \begin{bmatrix} x_I(t) \\ x_Q(t) \\ x_I^2(t)+x_Q^2(t) \end{bmatrix} + \frac{K_2}{4} \begin{bmatrix} 1 & -1 & 0 \\ 1 & 0 & -1 \end{bmatrix} \begin{bmatrix} G^2 \\ G^2 \\ G^2 \end{bmatrix} \tag{26}$$
$$= \frac{\sqrt{2}}{2} K_2 G \begin{bmatrix} x_I(t) \\ x_Q(t) \end{bmatrix}. \tag{27}$$

Clearly, the gain factors and angles cannot be set perfectly in practice. Consequently, the analog I/Q regeneration has its limits. In particular, if signals with high dynamic range are to be processed (e.g., multichannel mobile communications signals in base stations with a dynamic range of 80–100 dB), digital I/Q regeneration is the preferred method.

VII. Advantages of the Six-Port Technology: Applications

First, the basic characteristics of the architecture of Fig. 8 are summarized as follows.

- The six-port network is passive.
- The power detectors (square-law devices) are passive devices.
- G_i and φ_i do not need to have certain fixed values.

These characteristics lead to the fact that the realization of the six-port is rather uncomplicated given the fact that most parameter mismatches caused by the production are acceptable since they are dealt with in the I/Q regeneration procedure (of course, this is not true if an analog I/Q regeneration technique, as in Fig. 7, is employed). Hence, the performance of the six-port is not only a matter of the quality of the hardware, but it is, in particular, a matter of digital signal processing, which makes the six-port an initial candidate for software-defined radio. Software-defined radio receivers must be wide-band receivers in order to support as many services on different carrier frequencies as possible. Six-port receivers are wide-band receivers due to their passiveness [19]. Therefore, six-port technology is perfectly fitted to the requirements of wide-band software-defined radio [6].

Particularly interesting is the application of the six-port technology for very high frequencies where the whole six-port network can be realized as a very small microwave circuit using transmission-line elements.

However, the question arises of when the additional effort of the third ADC, filter, etc. is justified compared to a conventional direct down-conversion receiver. Of course, this depends on the application. The main reason for accepting the additional effort is when the six-port is the only technology that can serve certain requirements. With respect to the scope of this paper, these are clearly the requirements of a wide-band software-defined radio that is also capable of operating at very high frequencies. Several applications are planned to operate, e.g., in the 17-, 24-, or 60-GHz bands, particularly high data-rate applications such as future wireless local area networks (LANs) with data rates of up to 1 Gbit/s (see, e.g., [20]).

VIII. Conclusion

Based on the well-known background of the six-port technique, it has been shown that incident power observations provide signals that contain frequency-shifted versions of the input signal of the six-port (or five-port). Thus, it has been shown that the six-port (or five-port) is not a mystic component that provides (five or) six power measurements from which the in-phase and quadrature-phase components of the incoming bandpass signal can be calculated in an obscure way. There is rather an ordinary frequency conversion that takes place. To relate the six-port frequency-conversion structure to the well-known heterodyne and homodyne receiver structures, a systematic approach to frequency conversion has been given. From this systematic approach and from considerations on multiplicative and additive mixing, the super(position) homodyne receiver as a special case of the six-port (or five-port) receiver has been derived. The relationship between the classical six-port theory and the six-port as a means for frequency conversion has been established.

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