RF and IF Ports Matching Circuit Synthesis for a Simultaneous Conjugate-Matched Mixer Using Quasi-Linear Analysis

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***Abstract*—A quasi-linear two-port approach between RF and IF ports to design a simultaneous conjugate-matched mixer is presented in this paper. Conventionally, mixer design is treated as a nonlinear three-port device problem. Nonetheless, with the exception of the large-signal local oscillator (LO) that exists at the LO port, the input RF and output IF signals that exist at the RF and IF ports, respectively, are small signals. Consequently, mixers can be approximated as bilateral quasi-linear two-port circuits with a time-variant transfer function between the RF and IF ports, in which the LO port of the mixer is treated as part of the two-port network. With this approximation, it can be shown mathematically that the optimum source and load matching networks required for attaining simultaneous conjugate match at the RF and IF ports are actually time invariant, thus implying that it is possible to synthesize these optimum impedance values. This proposed mixer design technique, together with the equations derived, are verified with block-diagram simulation and experimental measurements of two 2.4-GHz RF/420-MHz IF double-balanced diode mixers.**

***Index Terms*—Circuit synthesis, linear circuits, mixers, nonlinear circuits, time-varying circuits.**

I. Introduction

In MIXER design, much effort has heretofore been directed to quantify the theoretical minimum loss of mixers [1], [2] and there is certainly no lack of papers on experimental results showing mixers with low conversion loss [3]–[5]. Many authors have also suggested ways to minimize mixer conversion loss using combinations of linear and nonlinear analysis methods [6], [7] and through proper termination of the idler frequencies at the RF and IF ports [8], [9] or via combined source–/load–pull techniques [10]. However, there is a dearth of papers on methods to properly terminate the RF and IF ports at the RF and IF frequencies, respectively, in order to achieve simultaneous conjugate match; a factor that is rather crucial in minimizing reflection losses at the RF and IF ports, which will, in turn, minimize the conversion loss of mixers. This is logical since mixers are inherently large-signal devices, which, by definition, are not possible to conjugate match [11]. In addition, being frequency-translating devices, the input RF and output IF signals of mixers are operating at different frequencies, which give rise to a time-dependent input–output relationship and, hence, make phase measurements using practical instruments impossible [12]. Nonetheless, the matching circuits for the mixer RF and IF ports can be designed by treating the mixer as a unilateral two-port device [13] or via harmonic-balance large-signal–small-signal analysis, which allows simultaneous conjugate match of the mixer RF and IF ports [11].

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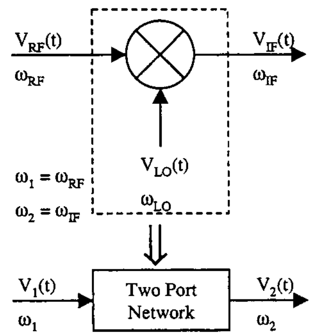


Fig. 1. Mixer as a quasi-linear time-variant two-port network.

Designing the matching circuits for the mixer RF and IF ports can be significantly simplified relative to the design method articulated in [11] and yet avoiding the oversimplifying unilateral assumption proposed in [13] when mixers are approximated as quasi-linear time-variant two-port circuits in which the local oscillator (LO) port of the original three-port mixer circuit is lumped as an integral part of the two-port network [3], [11], [13], as illustrated in Fig. 1. As a result, simultaneous conjugate match for the resultant bilateral two-port mixer network is achievable and it will be shown mathematically in this paper that the optimum source and load matching networks required at the RF and IF ports are actually time invariant. It is, therefore, possible to synthesize these optimum matching networks using practical matching circuits.

II. Mathematical Derivation

When modeling a mixer as a quasi-linear time-variant two-port network, the only difference with the more familiar linear time-invariant (LTI) two-port network is that the two ports are operating at different frequencies. The benefit of modeling the mixer as a quasi-linear two-port network is that it allows the application of the well-developed two-port network theory in its analysis. In Fig. 1, although ω_1 and ω_2 are equivalent to $\omega_{\rm RF}$ and $\omega_{\rm IF}$, respectively, there is a distinction between $v_1(t)$ and $v_{\rm RF}(t)$ or $v_2(t)$ and $v_{\rm IF}(t)$. The differences between these quantities is due to the inherent nonlinear nature of mixer circuits. When a signal is applied to the RF port of a mixer, a host of mixing sidebands are generated at both the RF and IF ports, albeit the original excitation signal consist of only a single sinusoid. In this paper, $v_{\mathrm{RF}}(t)$ and $v_{\mathrm{IF}}(t)$ represent the total voltage that exist at the RF and IF ports, respectively, whereas $v_1(t)$ represents the original single frequency excitation and $v_2(t)$ represents the desired mixing sideband, either up- or down-converted. The resultant unwanted mixing products that are generated in the frequency-translation process, especially the more important idler frequencies [8], [9], must be properly terminated. Ideally, the impedances presented to these unwanted mixing sidebands at the RF and IF ports should remain unchanged before and after the mixer is matched.

Referring to Fig. 1, let the input RF and output IF signals at ports 1 and 2 be

$$v_1(t) = V_1 \angle (\omega_1 t + \vartheta_1) \tag{1}$$

and

$$v_2(t) = V_2 \angle (\omega_2 t + \vartheta_2) \tag{2}$$

respectively. To show that the input–output transfer function is indeed that of a linear time-variant two-port network, (2) can be rewritten as

$$v_2(t) = h(t) \cdot v_1(t) \tag{3}$$

where

$$h(t) = \frac{V_2}{V_1} \angle (\vartheta_2 - \vartheta_1) \cdot 1 \angle \Delta\omega t \tag{4}$$

$$\Delta\omega = \omega_2 - \omega_1. \tag{5}$$

Equations (3)–(5) show that the input–output transfer function $h(t)$ is indeed that of a linear time-variant system [14]. It is composed of two parts, the former is merely that of the transfer function for an LTI system, while the latter has a time-dependent phase with constant unity magnitude. Had the two ports been operating at the same frequency, then this latter part of the transfer function would not have existed, thus relegating the system back to an LTI two-port network.

Having shown that the mixer transfer function is linear time variant, the next step is to use linear two-port network theory to show that the optimum impedances required at the RF and IF ports for attaining simultaneous conjugate match at both ports of the mixer are time invariant. The equations governing the relationship between the S-parameters and incident and reflected waves at ports 1 and 2 are shown as follows [15]:

$$b_1 = S_{11} a_1 + S_{12} a_2 \tag{6}$$

$$b_2 = S_{21} a_1 + S_{22} a_2 \tag{7}$$

where, for frequency translating devices, the incident and reflected waves at port 1, i.e., a_1 and b_1, are operating at angular frequency ω_1, while the incident and reflected waves at port 2, i.e., a_2 and b_2, are operating at angular frequency ω_2. Hence, the input and output reflection coefficients S_{11} and S_{22} are not time dependent since they are ratios of two waves operating at the same frequency and, thus, can be written as

$$S_{11} = |S_{11}| \angle \vartheta_{11} \tag{8}$$

$$S_{22} = |S_{22}| \angle \vartheta_{22} \tag{9}$$

which are no more than a pair of complex numbers similar to the time-invariant case. On the contrary, the forward and reverse transmission coefficients S_{21} and S_{12} are ratios of two waves operating at different frequencies and, hence, are time dependent. Consequently, these two quantities take the form of (4) and can be written as

$$S_{21} = |S_{21}| \angle \vartheta_{21} \cdot 1 \angle \Delta\omega t \tag{10}$$

$$S_{12} = |S_{12}| \angle \vartheta_{12} \cdot 1 \angle -\Delta\omega t \tag{11}$$

where the quantity $\Delta\omega$ is as defined in (5).

From [15], the source and load match required to simultaneous conjugate match a linear two-port network is

$$\Gamma_S = \frac{B_1 \pm \sqrt{B_1^2 - 4|C_1|^2}}{2C_1} \tag{12}$$

$$\Gamma_L = \frac{B_2 \pm \sqrt{B_2^2 - 4|C_2|^2}}{2C_2} \tag{13}$$

where

$$B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2 \tag{14}$$

$$B_2 = 1 + |S_{22}|^2 - |S_{11}|^2 - |\Delta|^2 \tag{15}$$

$$C_1 = S_{11} - \Delta S_{22}^* \tag{16}$$

$$C_2 = S_{22} - \Delta S_{11}^* \tag{17}$$

$$\Delta = S_{11} S_{22} - S_{21} S_{12} \tag{18}$$

in which S_{21} and S_{12} exist only in multiplied pair where their time-dependent phase terms cancel each other. This, in turn, infers that both Γ_S and Γ_L are not time-dependent quantities. As a result, when a mixer is modeled as a quasi-linear time-variant two-port network, the optimum matching networks required to simultaneous conjugate match its RF and IF ports are time invariant.

In general, mixers are divided into three categories depending on the relationship between the signals at the RF, IF, and LO ports [16]. The equations that describe the frequency relationship for the three classes of mixers are

$$f_{\mathrm{IF}} = f_{\mathrm{RF}} - f_{\mathrm{LO}} \tag{19}$$

$$f_{\mathrm{IF}} = f_{\mathrm{LO}} - f_{\mathrm{RF}} \tag{20}$$

$$f_{\mathrm{IF}} = f_{\mathrm{RF}} + f_{\mathrm{LO}}. \tag{21}$$

Equations (19) and (20) represent difference mixer with a low- and high-side LO, respectively, while (21) represents the sum mixer. It should be noted that the RF port is considered the input port, whereas the IF port is the output port irrespective of the mixer type even for the case of the sum mixer where the input RF frequency is lower than the output IF frequency; this is consistent with the definition used in [16]. Equations (8)–(11), though general they may seem, cannot be substituted directly into (12)–(18) for a difference mixer with a high-side LO. Sections III–V of this paper will detail this fact more thoroughly and subsequently derive an alternative set of equations that is applicable to this category of mixers.

III. DIFFERENCE MIXER WITH LOW-SIDE LO

When an RF voltage wave

$$v_1^+(t) = V_1 \angle (\omega_1 t + \vartheta_1) \tag{22}$$

is incident on the RF port of the mixer that has an applied LO voltage

$$v_{\mathrm{LO}}(t) = V_{\mathrm{LO}} \angle (\omega_{\mathrm{LO}} t + \vartheta_{\mathrm{LO}}) \tag{23}$$

then the transmitted IF voltage wave that emanates from the IF port [1] is

$$v_2^-(t) = K_{21} V_1 \angle \big((\omega_1 - \omega_{\mathrm{LO}}) t + \vartheta_1 - \vartheta_{\mathrm{LO}} \big) \tag{24}$$

and the forward transmission coefficient [15] is

$$S_{21} = K_{21} \angle (-\omega_{\mathrm{LO}} t - \vartheta_{\mathrm{LO}}). \tag{25}$$

The quantity K_{21} is the proportionality constant, which is equivalent to the conversion gain/loss of the mixer, and the term $-\omega_{\mathrm{LO}}$ in (25) is the angular frequency difference between the output and input ports (refer to (19)), which is equivalent to the term $\Delta\omega$ in (10). Similarly, it can be shown that S_{12} is

$$S_{12} = K_{12} \angle (\omega_{\mathrm{LO}} t + \vartheta_{\mathrm{LO}}). \tag{26}$$

Multiplying (25) and (26) together yields

$$S_{21} S_{12} = K_{21} K_{12} \angle 0^\circ \tag{27}$$

which evidently does not consist of any time-dependent term; this serves to reinforce the argument made in the previous section that Γ_S and Γ_L are not time-dependent quantities.

Having all the necessary equations derived, the succeeding step is to validate these equations with computer-aided design (CAD) simulation using ideal blocks; the simulator used for this purpose is HP-MDS. A hard fact pointed out by (10) and (11) is that S_{21} and S_{12} are time-dependent quantities and, therefore, must be extracted at a similar instant of time. This, of course, precludes the use of any practical measurement equipment, thus making simulation the only design option available. Through the use of a simulator, a snapshot of time can be taken, as illustrated in Fig. 2, where, in the harmonic-balance simulation, the phase of the LO signal must be set to similar values for the forward and reverse S-parameters extraction. This is to ensure that the contribution of ϑ_{LO} in the extracted S_{21} and S_{12} is of equal value, thereby resulting in exact cancellation of the ϑ_{LO} term when these two quantities are multiplied together, as dictated by (25)–(27). For convenience, they can simply be set to zero during the simulation. Post-processing is then used to select the appropriate mixing sidebands for calculating the S-parameters, where the simulated S_{21} and S_{12} are merely a set of complex numbers with no time dependence because they are extracted at a particular instant of time. The basic mixer topology used in the simulation in Fig. 2 is a 2.4-GHz RF/420-MHz IF double-balance diode mixer with a low-side LO diagrammed in Fig. 3. All the component models used in the simulation are ideal, except the diode ring, which is from Agilent Technologies, Palo Alto, CA (P/N 5082-2830).

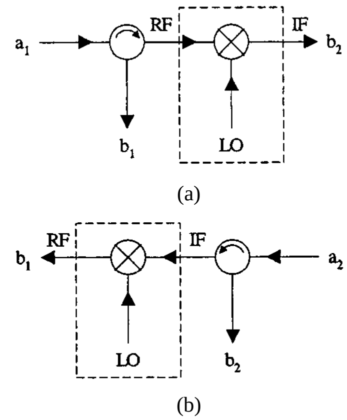


Fig. 2. (a) Forward and (b) reverse S-parameters extraction.

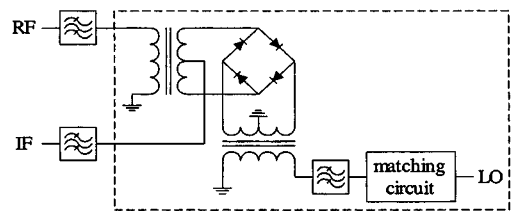


Fig. 3. Basic two-port quasi-linear mixer.

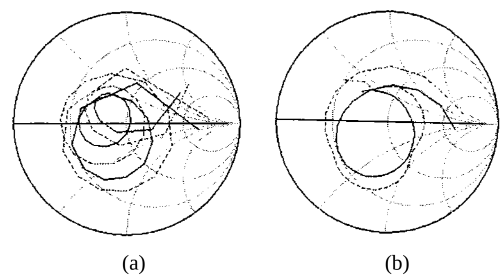


Fig. 4. Simulated: (a) input and (b) output reflection coefficients before matching. (a) S_{11} - - -. Γ_S^* —. (b) S_{22} - - -. Γ_L^* —.

Fig. 4 depicts the simulated input and output reflection coefficients of the mixer for a 400-MHz bandwidth centered at 2.4-GHz RF and 420-MHz IF with the LO signal fixed at 1.98 GHz. The traces for S_{11} (or S_{22}) and Γ_S^* (or Γ_L^*) do not overlap; the simple reason being mixers, similar to small-signal amplifiers, are bilateral devices. If the mixer is treated as a unilateral device when designing the RF and IF matching circuits, then the final designed circuit would have been nonoptimal with power being reflected at the input RF and output IF ports, thereby resulting in higher insertion loss.

The extracted S-parameters can be substituted directly into (12)–(18) to obtain Γ_S and Γ_L; the matching circuits are then realized using two-port S-parameter networks available in the simulator. These ideal matching networks are subsequently cascaded to the RF and IF ports of the basic mixer circuit in Fig. 3 and the entire circuit is resimulated to verify its performance, which is tabulated in Table I. Input and output return losses are better than 70 dB, while the conversion gain, neglecting mathematical truncation, is virtually equivalent to the maximum transducer gain (Gmax) [15] predicted before the mixer was matched. In the simulation, the matching circuits are implemented using mathematical S-parameter blocks, which

TABLE I
SIMULATED MIXER PERFORMANCE AFTER MATCHING

RF/GHz	IF/MHz	S_{11}/dB	S_{22}/dB	S_{21}/dB	Gmax
2.20	220	-76.58	-79.69	-4.06	-4.06
2.30	320	-88.44	-83.73	-4.03	-4.03
2.40	420	-79.43	-78.56	-4.02	-4.03
2.50	520	-82.75	-81.32	-4.03	-4.03
2.60	620	-76.99	-72.80	-4.04	-4.05

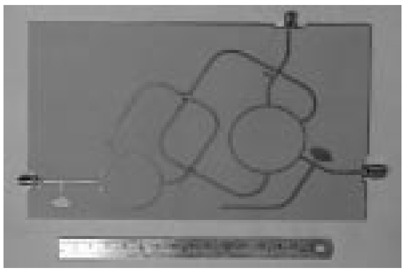
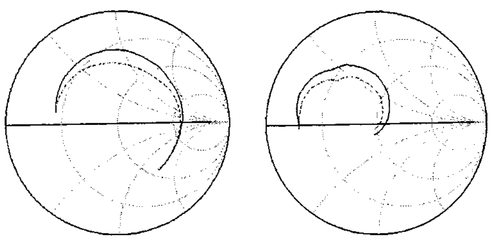
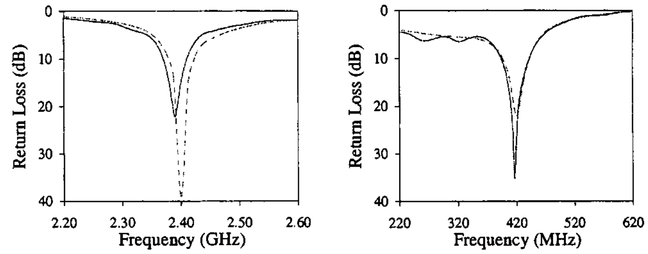


Fig. 5. Photograph of actual mixer circuit.



(a) (b)

Fig. 6. Simulated: (a) input and (b) output reflection coefficients before matching. (a) S_{11} - - -. Γ_S^* —. (b) S_{22} - - -. Γ_L^* —.



(a) (b)

Fig. 7. Simulated and measured: (a) input and (b) output return losses after matching. (a) Simulated S_{11} - - -. Measured S_{11} —. (b) Simulated S_{22} - - -. Measured S_{22} —.

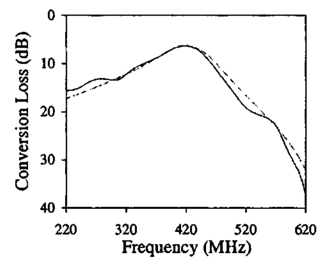


Fig. 8. Simulated and measured conversion losses after matching. Simulated S_{21} - - -. Measured S_{21} —.

TABLE II
SIMULATED AND MEASURED PERFORMANCE AFTER MATCHING

	Simulated / f_o	Measured / f_{offset}
S_{11}	-39.07dB / 2.40GHz	-22.21dB / 2.39GHz
S_{22}	-22.04dB / 420MHz	-35.08dB / 416MHz
S_{21}	-6.29dB / 420MHz	-6.37dB / 418MHz

means that they are not bounded by the bandwidth-limiting Bode–Fano criteria applicable when physical matching circuits are used [15]. This is the reason why the input and output return losses are excellent over the entire 400-MHz bandwidth.

To further verify that the equations and design methodologies described above are applicable to practical mixer circuits, a 2.4-GHz RF/420-MHz IF double-balanced diode ring mixer is designed following the above-mentioned design procedure. The nonlinear element employed is the HP ring diode used in the previous example, while the rest of the circuit is implemented using microstrip techniques fabricated on an RO4003 substrate. Fig. 5 shows a photograph of the actual mixer circuit. The RF and IF ports are at the left- and right-hand sides of the printed circuit board (PCB), respectively, while the LO port is at the top side; matching for all the ports are accomplished via narrow-band single-stub match using the design and optimization procedure articulated in [17]. The two circular structures on the PCB are rat-race hybrids functioning as the 2.4-GHz RF and 1.98-GHz LO balun.

The simulated input and output reflection coefficients of this mixer without the matching circuits are depicted in Fig. 6 and the predicted maximum transducer gain is −6.05 dB. The simulated and measured return losses of this mixer and its conversion losses with the RF and IF matching circuits in place are diagrammed in Figs. 7 and 8. As tabulated in Table II, the measured performance is slightly offset from the designed frequency. Nevertheless, the frequency offset is relatively small and the −6.29 dB simulated and −6.37 dB measured conversion gain of this mixer at the design frequency of 2.4-GHz RF/420-MHz IF are close to the −6.05 dB maximum transducer gain predicted before the circuit was matched.

IV. DIFFERENCE MIXER WITH HIGH-SIDE LO

By following the argument used in Section III, it can be shown that when a voltage wave that takes the form of (22) is incident on the RF port of a difference mixer with a high-side LO that has an LO voltage of the form of (23), the transmitted IF voltage wave that emanates from the IF port is

$$v_2^-(t) = K_{21} V_1 \angle\big((\omega_{\text{LO}} - \omega_1)\, t + \vartheta_{\text{LO}} - \vartheta_1\big) \tag{28}$$

and S_{21} is

$$S_{21} = K_{21} \angle\big((\omega_{\text{LO}} - 2\omega_1)\, t + \vartheta_{\text{LO}} - 2\vartheta_1\big). \tag{29}$$

Likewise, it can be shown that S_{12} is

$$S_{12} = K_{12} \angle\big((\omega_{\text{LO}} - 2\omega_2)\, t + \vartheta_{\text{LO}} - 2\vartheta_2\big). \tag{30}$$

Multiplying (29) and (30) together and with the help of (20) results in

$$S_{21} S_{12} = K_{21} K_{12} \angle 2\left(\vartheta_{\text{LO}} - \vartheta_1 - \vartheta_2\right) \tag{31}$$

which is independent of time, but a function of the RF, IF, and LO phase; this, in turn, implies that the optimum source and load match that need to be presented to the RF and IF ports of the mixer are dependent on these three quantities. If this statement is true, then it would be impossible to simultaneous conjugate match this type of mixers. Fortunately, this is not the case and it can be proven that a solution comparable to that of (27), which is applicable to a difference mixer with a low-side LO, also exist for a difference mixer with a high-side LO.

For clarity and easy explanation, (28) is rewritten as

$$v_2^-(t) = K_{21}\angle(\omega_{\rm LO}t+\vartheta_{\rm LO})\cdot V_1\angle-(\omega_1 t+\vartheta_1) \quad (32)$$

which shows that when the input RF signal is down-converted to the IF signal, it experiences amplification or attenuation depending on the term K_{21}, a time-dependent phase shift introduced by $\omega_{\rm LO}t+\vartheta_{\rm LO}$ and a phase inversion due to $-(\omega_1 t+\vartheta_1)$. With this in mind, (29) and (32) can be further rewritten as

$$v_2^-(t) = S'_{21}v_1^+(t)^* \quad (33)$$

$$S_{21} = S'_{21}\cdot 1\angle-2\vartheta'_1 \quad (34)$$

where

$$S'_{21} = K_{21}\angle(\omega_{\rm LO}t+\vartheta_{\rm LO}) \quad (35)$$

$$\vartheta'_1 = \omega_1 t+\vartheta_1. \quad (36)$$

Correspondingly, the S_{12} in (30) can also be rewritten as

$$S_{12} = S'_{12}\cdot 1\angle-2\vartheta'_2 \quad (37)$$

where

$$S'_{12} = K_{12}\angle(\omega_{\rm LO}t+\vartheta_{\rm LO}) \quad (38)$$

$$\vartheta'_2 = \omega_2 t+\vartheta_2. \quad (39)$$

With this, the RF input impedance of the mixer when the IF port is terminated with Γ_L at the IF frequency and the IF output impedance of the mixer when the RF port is terminated with Γ_S at the RF frequency are

$$\Gamma_{\rm in} = S_{11}+\frac{S'_{21}{}^*S'_{12}\Gamma_L^*}{1-S_{22}{}^*\Gamma_L^*} \quad (40)$$

$$\Gamma_{\rm out} = S_{22}+\frac{S'_{21}S'_{12}{}^*\Gamma_S^*}{1-S_{11}{}^*\Gamma_S^*} \quad (41)$$

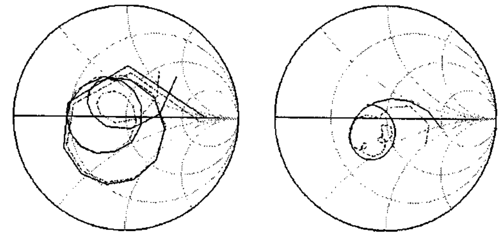
where both the input and output impedances are time invariant since the time-dependent phase of both S'_{21} and S'_{12} cancel each other due to the complex conjugate function performed on either one of these quantities. On top of that, $\Gamma_{\rm in}$ and $\Gamma_{\rm out}$ are not dependent on the phase of the input RF, output IF, and LO signals, contrary to what had been demonstrated earlier in (31).

To simultaneous conjugate match a difference mixer with a high-side LO, the following modifications will have to be made to C_1, C_2, and Δ in (16)–(18)

$$C_1 = S_{11}-\Delta S_{22} \quad (42)$$

$$C_2 = S_{22}-\Delta^* S_{11} \quad (43)$$

$$\Delta = S_{11}S_{22}^*-S'_{21}{}^*S'_{12} \quad (44)$$



(a) (b)

Fig. 9. Simulated: (a) input and (b) output reflection coefficients before matching. (a) S_{11} - - -. Γ_S^* —. (b) S_{22} - - -. Γ_L^* —.

TABLE III
SIMULATED MIXER PERFORMANCE AFTER MATCHING

RF/GHz	IF/MHz	S_{11}/dB	S_{22}/dB	S_{21}/dB	Gmax
2.60	220	-63.90	-64.15	-3.36	-3.35
2.50	320	-82.67	-69.17	-3.41	-3.40
2.40	420	-72.46	-65.51	-3.45	-3.45
2.30	520	-70.65	-68.31	-3.49	-3.49
2.20	620	-69.41	-65.60	-3.53	-3.53

where S'_{21} and S'_{12} exist only in (44) in mutually multiplied pairs with S'_{21} in complex conjugate form, thus resulting in

$$S'_{21}{}^*S'_{12} = K_{21}K_{12}\angle 0^\circ \quad (45)$$

which is basically a direct equivalent of (27) for a difference mixer with a low-side LO.

To attest the validity of the equations derived, CAD tools are utilized to simulate a 2.4-GHz RF/420-MHz IF double-balanced difference mixer with a high-side LO using ideal blocks. The harmonic-balance simulation setup for extracting the S-parameters of the mixer circuit is analogous to that for the difference mixer with a low-side LO, as illustrated in Fig. 2. A fundamental aspect like setting the phase of the LO signal to similar values during the forward and reverse S-parameters extraction is still necessary since the forward and reverse transmission coefficients S'_{21} and S'_{12} are still dependent on the phase of the LO signal. For convenience, the phase of the incident signals at the RF or IF ports were set to zero. By doing so, the extracted forward and reverse transmission coefficients directly yield S'_{21} and S'_{12} in (35) and (38), parameters which are devoid of the phase angle contributed by the incident waves, instead of S_{21} in (34) and S_{12} in (37). As such, the simulated S-parameters can be substituted into (12)–(15) and (42)–(44) directly to obtain the necessary source and load match needed to simultaneous conjugate match the mixer.

Fig. 9 shows the simulated input and output reflection coefficients of the mixer diagrammed in Fig. 3 for a 400-MHz bandwidth centered at 2.4-GHz RF and 420-MHz IF with the LO signal fixed at 2.82 GHz. Neither the traces for S_{11} and Γ_S^*, nor S_{22} and Γ_L^* overlap.

The simulated performance of this mixer with the matching circuits in place is listed in Table III. Input and output return losses are better that 60 dB, while the conversion gain, neglecting mathematical truncation, is virtually equivalent to the maximum transducer gain predicted before the mixer was matched. At first sight, the better than −4 dB Gmax listed may seem contrary to what is theoretically achievable with a

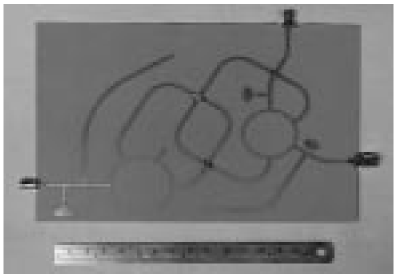
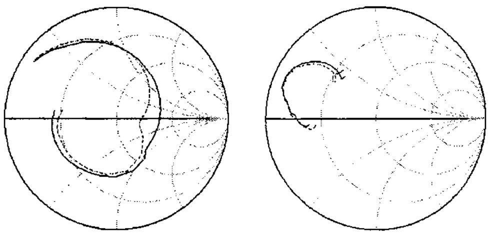
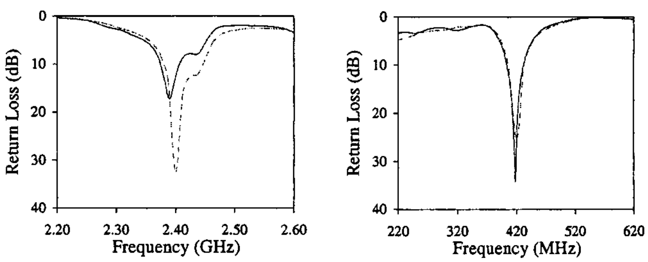


Fig. 10. Photograph of actual mixer circuit.



(a) (b)

Fig. 11. Simulated: (a) input and (b) output reflection coefficients before matching. (a) S_{11} - - -. Γ_S^* —. (b) S_{22} - - -. Γ_L^* —.



(a) (b)

Fig. 12. Simulated and measured: (a) input and (b) output return losses after matching. (a) Simulated S_{11} - - -. Measured S_{11} —. (b) Simulated S_{22} - - -. Measured S_{22} —.

double-balanced diode mixer [1]. However, this should not be surprising since the authors of [8] and [9] were able to obtain simulated conversion loss as low as 3.6 and 3.9 dB, respectively, by optimally terminating the important idler frequencies. It is, nonetheless, not the objective of this paper to illustrate this concept. The exceptionally promising attainable conversion loss is obtained as a matter of coincidence; there was no deliberate effort to establish the optimum loading conditions for the various idler frequencies.

To further validate the applicability of the equations and design methodologies on a practical difference mixer with a high-side LO, a 2.4-GHz RF/420-MHz IF double-balanced diode ring mixer is designed following the aforementioned design procedure. Fig. 10 shows a photograph of this mixer circuit. It is very similar to the difference mixer with a low-side LO depicted in Fig. 5, except that the LO frequency is at 2.82 GHz instead of 1.98 GHz.

The simulated input and output reflection coefficients of this mixer without the matching circuits are depicted in Fig. 11, which has a maximum transducer gain of −6.45 dB when it is simultaneous conjugate matched at 2.4-GHz RF/420-MHz IF. The simulated and measured return losses of this mixer with the RF and IF matching circuits in place are diagrammed in Fig. 12,

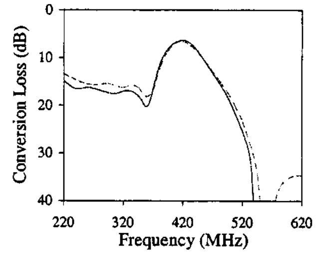


Fig. 13. Simulated and measured conversion losses after matching. Simulated S_{21} - - -. Measured S_{21} - - -.

TABLE IV
SIMULATED AND MEASURED PERFORMANCE AFTER MATCHING

	Simulated / f_o	Measured / f_{offset}
S_{11}	-32.53dB / 2.40GHz	-17.30dB / 2.39GHz
S_{22}	-25.32dB / 420MHz	-34.17dB / 418MHz
S_{21}	-6.62dB / 420MHz	-6.36dB / 418MHz

while Fig. 13 shows the mixer conversion loss. As tabulated in Table IV, the measured performance is slightly offset from the designed frequency and, in Fig. 13, the measured conversion loss data is only plotted up to 540 MHz because, toward the higher end of the IF frequency range, the insertion loss is very high (predicted by the simulated results) and since the HP8753E network analyzer used to make this measurement has limited dynamic range at its R channel input, it is unable to phase lock to the weak IF signal present at its R channel input. Consequently, that part of the measurement results are not valid and, therefore, not shown. Excluding these anomalies, the simulated and measured traces have almost identical profile, though not entirely overlapping each other. Additionally, the −6.62 dB simulated and −6.36 dB measured conversion gain of this mixer at the design frequency of 2.4-GHz RF/420-MHz IF are close to the −6.45 dB maximum transducer gain that was predicted before the mixer was matched.

V. Sum Mixer

The last type of mixers are the sum mixers with their frequency relationship governed by (21). If the RF and IF subscripts in this equation are interchanged, the result will be (19). This, in turn, implies that, mathematically, sum mixers are analogous to difference mixers with a low-side LO with their input and output ports interchanged. As a result, the source and load match for sum mixers can be found using (12)–(18), similar to difference mixers with a low-side LO.

VI. Conclusion

It has been shown that, by modeling mixers as quasi-linear time-variant two-port networks, in which the LO port is included as part of the two-port network, it is mathematically possible to synthesize the matching circuits to simultaneous conjugate match the mixer input RF and output IF ports. In addition, mathematical derivation reveals that these required optimum source and load impedances are time invariant, which implies that it is possible to realize these optimum impedance values using practical matching circuits. With the newly derived equations, a synthesis instead of analysis approach is feasible when designing matching circuits for mixers. The design techniques involved are used prevalently in linear circuit design and are readily amenable to CAD techniques.

Had it been achievable to realize frequency translation without resorting to nonlinear circuits, the derived equations would have been mathematically exact. Nevertheless, this is not the case and, therefore, the other unwanted mixing products that are generated in the frequency-mixing process, especially the more important idler frequencies [8], [9], must be properly terminated; ideally, the impedances presented to these unwanted mixing products at the RF and IF ports should remain unchanged before and after the mixer is matched. Nonetheless, a combination of the design methods articulated in [8] and [9] can be used in conjunction with the equations described in this paper, whereby the important idler frequencies are optimally terminated to obtain the highest possible maximum transducer gain before the mixer is simultaneous conjugate matched, thus ensuring the final mixer circuit has the lowest conversion loss/highest conversion gain while its RF and IF ports are simultaneous conjugate matched. However, using this combined design technique, the matching circuits required are rather difficult, though not impossible, to realize using practical circuits and, hence, are neither considered, nor implemented in this paper.

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