

High Dynamic Range Double-Balanced Active Mixers Using Lossless Feedback

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Abstract - A further development in the application of negative feedback techniques to active double-balanced mixers is described, in which the concept of single-transformer lossless feedback is employed to improve both the intermodulation intercept point and the noise figure. A generalized topology is described, and test data describing and comparing the performance of example series-shunt and lossless feedback active mixers is presented.

I. INTRODUCTION

With the rapid expansion of wireless communications and the subsequent crowding of the allotted RF spectrum, designers of communications systems have been heavily tasked to determine ways to improve the dynamic range of receiver front ends and at the same time decreasing the power supply demands in portable communications equipment. While numerous solutions exist for the design of low-noise amplifiers with high intermodulation performance, the mixer remains the weak link in receiver system design.

Recently, a feedback topology used in the linearization of small-signal amplifiers was employed for linearizing double-balanced active mixers [1, 2, 3]. This method provides a high degree of improvement in intermodulation distortion (IMD), but does not improve the noise figure (NF). Subsequently, an alternate feedback topology has been adapted for the improvement in the dynamic range of the double-balanced mixer both in terms of IMD and NF, and a portion of the results are presented herein.

II. SERIES-SHUNT FEEDBACK ACTIVE MIXERS

The double-balanced active mixer, originally invented by Howard Jones in 1966 [4], has undergone a wide variety of modifications and adaptations, but the linearization of this most basic circuit has continued to elude designers. By making use of a transistor amplifier feedback technique commonly referred to as series-shunt [5], it has become possible to improve the IMD performance of the double-balanced active mixer in the same manner as for the amplifier. Fig. 1 describes a series-shunt feedback mixer in its basic functional form. Transistors Q3 and Q6 convert an input differential IF (or RF) signal ω_s into a pair of differential currents, which are subsequently switched by two differential transistor pairs Q1/Q2 and Q4/Q5.

at a rate determined by a local oscillator signal ω_L , resulting in the following four signal voltages at the collectors of transistors Q1, Q2, Q4, and Q5, respectively:

$$v_{C1} = + \frac{A \times A_V \times [\text{Cos}(\omega_L - \omega_s) t + \text{Cos}(\omega_L + \omega_s) t]}{2} \quad (1)$$

$$v_{C2} = - \frac{A \times A_V \times [\text{Cos}(\omega_L - \omega_s) t + \text{Cos}(\omega_L + \omega_s) t]}{2} \quad (2)$$

$$v_{C4} = + \frac{A \times A_V \times [\text{Cos}(\omega_L - \omega_s) t + \text{Cos}(\omega_L + \omega_s) t]}{2} \quad (3)$$

$$v_{C5} = - \frac{A \times A_V \times [\text{Cos}(\omega_L - \omega_s) t + \text{Cos}(\omega_L + \omega_s) t]}{2} \quad (4)$$

where A is the peak signal voltage and A_V is the voltage gain which is determined by:

$$A_V = \sqrt{\frac{R_I}{R_2 + r_e}} \quad (5)$$

The second term in the denominator of (5) represents the nonlinear emitter resistance of the amplifying transistors Q3 and Q6, which is the primary source of IMD in the double-

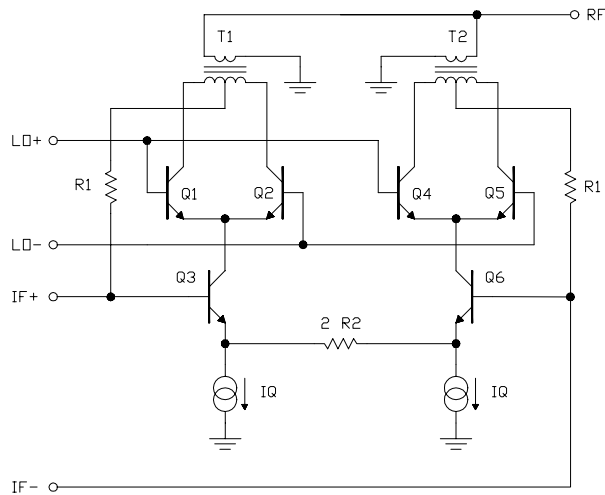


Fig. 1 - Series-Shunt Feedback Double-Balanced Active Mixer

balanced active mixer. These four signals are combined by hybrid transformers T_1 and T_2 [6, 7], resulting in a pair of feedback signal voltages:

$$v_{fb1} = -A \times A_V \times \cos \omega_s t \quad (6)$$

$$v_{fb2} = +A \times A_V \times \cos \omega_s t \quad (7)$$

and an output signal voltage

$$v_{out} = A \times A_V \times [\cos(\omega_L - \omega_s)t + \cos(\omega_L + \omega_s)t] \quad (8)$$

The hybrid transformers also serve to cancel the LO signal voltage at the collectors of switching transistors Q_1 , Q_2 , Q_4 , and Q_5 , thus further improving the compression characteristics of the mixer. Other parameter of interest include the input and output impedances, which, if the turns ratio of the hybrid transformers is 1:1:1, are determined by:

$$R_{in} = R_{out} = 2 \times R_L = \sqrt{R_L \times (R_2 + r_e)} \quad (9)$$

This mixer is relatively simple to make, and a similar version which utilizes a resistor network in place of the hybrid transformer [1] has comparable performance [2] and is practical for MMIC applications. The series-shunt feedback mixer has been demonstrated to have superior IMD qualities over the traditional double-balanced active mixer [2]. It does not, however, offer any improvement in NF.

III. LOSSLESS FEEDBACK ACTIVE MIXERS

In the design of small-signal amplifiers, the lossless feedback topology stands out as being the most effective and economical means of obtaining good NF and IMD performance [8, 9, 10]. Fig. 2 illustrates an adaptation of the single-transformer lossless feedback amplifier topology as a means of improving the NF and IMD performance of a double-balanced active mixer [11]. Here, the input IF (or RF) signal ω_s is applied differentially to a pair of lossless feedback transformers T_3 and T_4 , which results in a pair of differential input currents

$$I_{in+} = I_Q + \frac{A \times \cos \omega_s t}{R_{in}} \quad (10)$$

$$I_{in-} = I_Q - \frac{A \times \cos \omega_s t}{R_{in}} \quad (11)$$

The input resistance R_{in} is determined by:

$$R_{in} = \frac{2 \times K^2 \times R_L \times (M + N + 1)}{M^2 \times (M + N)} \quad (12)$$

where K is the turns ratio of the hybrid transformers T_1 and T_2 , M and N are the output winding turns ratios of the feedback transformers T_3 and T_4 , and R_L is the output RF (or IF) load resistance. The amplifying transistors Q_3 and Q_6 pass these

input currents to the differential switching transistor pairs Q_1/Q_2 and Q_4/Q_5 , respectively, resulting in two pairs of differential currents:

$$I_{C1} = I_{in+} \times \frac{1 - \cos \omega_L t}{2} = \frac{I_Q \times (1 - \cos \omega_L t)}{2} + \quad (13)$$

$$+ \frac{A \times \left[\cos \omega_s t - \frac{\cos(\omega_s - \omega_L)t + \cos(\omega_s + \omega_L)t}{2} \right]}{2 \times R_{in}}$$

$$I_{C2} = I_{in+} \times \frac{1 + \cos \omega_L t}{2} = \frac{I_Q \times (1 + \cos \omega_L t)}{2} + \quad (14)$$

$$+ \frac{A \times \left[\cos \omega_s t + \frac{\cos(\omega_s - \omega_L)t + \cos(\omega_s + \omega_L)t}{2} \right]}{2 \times R_{in}}$$

$$I_{C4} = I_{in-} \times \frac{1 - \cos \omega_L t}{2} = \frac{I_Q \times (1 - \cos \omega_L t)}{2} - \quad (15)$$

$$- \frac{A \times \left[\cos \omega_s t + \frac{\cos(\omega_s - \omega_L)t + \cos(\omega_s + \omega_L)t}{2} \right]}{2 \times R_{in}}$$

$$I_{C5} = I_{in-} \times \frac{1 + \cos \omega_L t}{2} = \frac{I_Q \times (1 + \cos \omega_L t)}{2} - \quad (16)$$

$$- \frac{A \times \left[\cos \omega_s t - \frac{\cos(\omega_s - \omega_L)t + \cos(\omega_s + \omega_L)t}{2} \right]}{2 \times R_{in}}$$

These two pairs of differential currents are then combined by hybrid transformers T_1 and T_2 [6, 7], which results in an RF

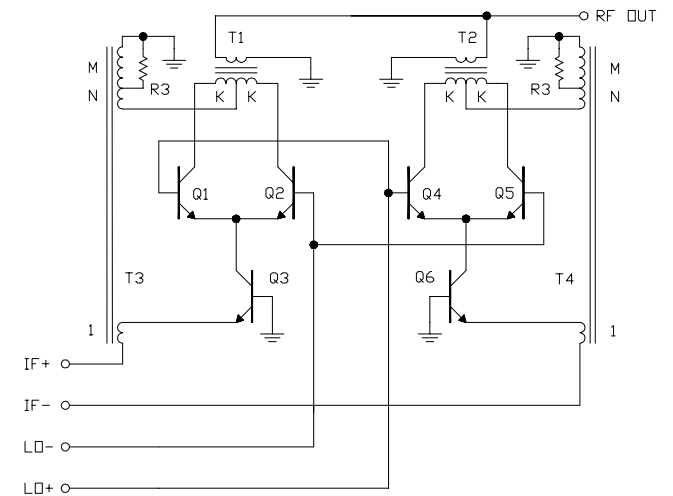


Fig. 2 - Lossless Feedback Double-Balanced Active Mixer

(or IF) output current and voltage

$$i_{out} = K \times (I_{C1} - I_{C2}) - K \times (I_{C4} - I_{C5}) = \quad (17)$$

$$= 2 \times A \times K^2 \times \frac{\text{Cos}(\omega_s - \omega_L) t + \text{Cos}(\omega_s + \omega_L) t}{R_{in}}$$

$$v_{out} = 2 \times A \times K^2 \times R_L \times \quad (18)$$

$$\times \frac{\text{Cos}(\omega_s - \omega_L) t + \text{Cos}(\omega_s + \omega_L) t}{R_{in}}$$

and a differential pair of feedback currents

$$I_{FB1} = I_{C1} + I_{C2} = I_Q + \frac{A \times \text{Cos} \omega_s t}{R_{in}} \quad (19)$$

$$I_{FB2} = I_{C4} + I_{C5} = I_Q - \frac{A \times \text{Cos} \omega_s t}{R_{in}} \quad (20)$$

which are then conducted to the output windings of the feedback transformers T_3 and T_4 , respectively, thus forming the negative feedback path necessary for linearization. The IF (or RF) terminating resistance R_3 is determined by:

$$R_3 = \frac{2 \times K^2 \times R_L}{M + N} \quad (21)$$

which represents a unique quality of the lossless feedback mixer, wherein the input has a wideband resistive input impedance which is insensitive to the IF load impedance.

IV. INTERMODULATION AND NOISE SOURCES

The degenerated common-emitter driver of the series-shunt feedback mixer and the common-base driver of the lossless feedback mixer have distinctly different characteristics with regard to intermodulation and noise. First, distortion in an undegenerated common-emitter driver tends to decrease with frequency, whereas distortion in the common-base driver rises monotonically with increased ω , and is independent of $\Delta\omega$ [12]. The common-base driver is preferred when wideband operation is required as it provides a nearly constant input impedance and gain [12]. The NF of common-base drivers increases monotonically with increasing bias current and in general tends to be higher than that of common-emitter drivers [12].

If the gain and output noise power of the driver stages is constant across all frequencies, the square-wave switching process would increase the input-referred noise contribution from the driver stage by a factor of $(\pi/2)^2$, or 3.9dB. This is a result of the square-wave LO mixing noise at various frequencies down to the IF, and on a linear scale the overall noise power of the mixer would be:

$$NF = N_D \times \left(\frac{\pi}{2}\right)^2 + N_{sw} \quad (22)$$

where N_{sw} is the noise contribution from the differential switching pairs Q_1/Q_2 and Q_4/Q_5 , and N_D is the input-referred noise

of the driver stages Q_3 and Q_6 [13, 14], consisting of base shot noise (N_b), collector shot noise (N_c), and thermal noise (N_t) [13, 14]:

$$N_D = 1 + N_c + N_b + N_t \quad (23)$$

With respect to the series-shunt feedback mixer, we can expect that the NF will be somewhat greater than we would determine from (22) as there are the additional thermal noise contributions from the feedback resistances R_1 and R_2 , as well as contributions from the biasing current sources. Conversely, we would expect that the NF of the lossless feedback mixer to be considerably closer to (22) given that the feedback transformers will ideally contribute very little to the overall NF. In either case, (22) shows that even with the best of transistors the NF of double-balanced active mixers will remain high.

V. EXPERIMENTAL RESULTS

For the purposes of demonstration, representative series-shunt feedback and lossless feedback mixers having the following parameters were constructed and evaluated:

Transistors:	Intersil (née Harris, née RCA) CA3102
Biasing:	$I_Q = 10\text{mA}$ $V_{B3, B6} = 2.65\text{V}$ $V_{B1, B2, B4, B5} = 5.00\text{V}$ $V_{CC} = 12.00\text{V}$
Impedance:	50 ohms (input and output)
LO signal:	10MHz, 0dBm
IF signals:	950kHz and 1100kHz

For the series-shunt feedback mixer, the particulars are:

T_1, T_2 :	MiniCircuits T4-1 (1:1:1)
R_1 :	330
R_2 :	33

and for the lossless feedback mixer:

T_1, T_2 :	MiniCircuits T4-1 (1:1:1)
T_3, T_4 :	MiniCircuits T2-613-1 (1:1:2)
R_3 :	30

In both cases, appropriate input transformers (MiniCircuits T4-1) are used to supply differential IF and LO input signals, commercial parts being used throughout in order that the circuits be repeatable. The test results are shown in Figures 3 and 4, as well as being summarized in Table 1. The series-shunt feedback mixer shows reasonable good linearity, and the results obtained here are consistent with those measured earlier for this type [2]. Compression takes place abruptly, which is characteristic of feedback circuits. The intermodulation products remain at 30dBc or better below the point of compression.

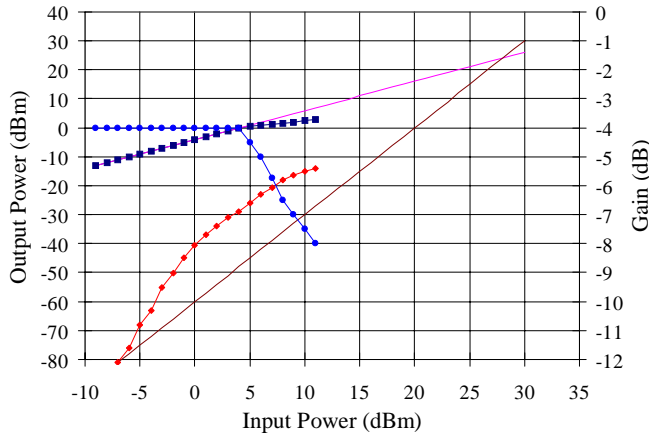


Figure 3 Type 2 Series-Shunt Feedback Mixer

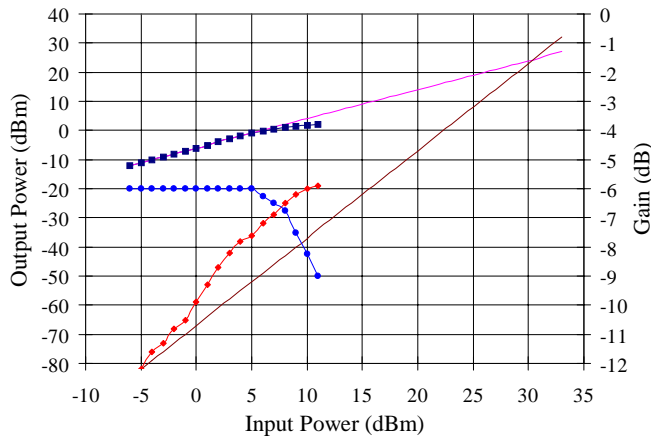


Figure 4 Lossless Feedback Mixer

Intermodulation ratio (IMR) expansion begins early, indicating that the series-shunt topology has a moderate ability to compensate for the initial onset of compression.

For the lossless feedback active mixer, the test results shown in Figure 4 indicate that there has been an improvement of as much as 20dB in the IM products, with the IM products remaining at 30dBc or better even slightly beyond the onset of compression. IMR expansion is delayed an additional 8dB or so beyond the point experienced with the series-shunt mixer, and the degree of expansion beyond this point is less pronounced. In general, these differences are typical of

TABLE I
FEEDBACK MIXER PERFORMANCE CHARACTERISTICS

Configuration	IIP ₃	P _{1dB}	Gain	NF
Gilbert Cell	+15.75dBm	+4.5dBm	-2.5dB	11.89dB
Type 2 Series-Shunt	+28dBm	+4dBm	-4dB	14.38dB
Lossless Feedback	+30.3dBm	+3dBm	-6dB	13.80dB

the performance to be seen in series-shunt and lossless feedback amplifiers.

V. PERFORMANCE COMPARISON

Table 1 summarizes the input intermodulation intercept point (IIP₃), the 1dB compression point (P_{1dB}), gain, and NF performance for the described circuits as well as for a comparable Gilbert Cell mixer. It can be readily seen that the lossless feedback mixer is an improvement over the earlier series-shunt feedback mixer in terms of both IIP₃ and NF. The variations in NF performance between the three example circuits is consistent with established theory relating to the effects of applying feedback to noisy two-ports [15].

VI. CONCLUSIONS

It has been demonstrated that the dynamic range of active mixers can be improved by applying feedback techniques commonly employed in small signal amplifiers. Test results shown here indicate that the application of lossless feedback results in a moderate improvement in IMD and NF performance over that of the previously described series-shunt feedback mixer, and the substantial improvement in dynamic range over the more common Gilbert Cell mixer is advanced.

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