A LINEARIZED ACTIVE MIXER

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ABSTRACT

A method is described in which the intermodulation performance of an active mixer is improved by recovering the amplified output IF signal and combining it with the input IF signal in a pair of embedded series/shunt feedback amplifiers, thereby overcoming the nonlinearities of the active devices. Performance characteristics are compared with a previous realization[1,2], as well as with diode ring, switching MOSFET, and transistor tree mixers.

INTRODUCTION

Mixers and modulators are essential building blocks in RF communications systems. Various realizations employing diodes, switching MOSFETs, dual-gate FETs, and the highly popular transistor tree (aka Gilbert Cell) mixer provide the necessary means by which frequency conversion, modulation, or demod-ulation takes place. In all realizations, the nonlinearities of the devices, either directly or indirectly, cause distortion of the desired signals when two or more of the signals interact, a phenomenon know to the profession as intermodulation distortion (IMD).

Much has been written in the professional literature with regard to the sources of IMD, and a furtherance of that discussion is not within the scope of this paper. Rather, a brief discussion will be presented with regard to two of the more common mixer realizations, diode rings and transistor trees, where this characteristic will be examined for later comparison with the recently introduced feedback mixer, in which the integrity of the incoming intermediate frequency (IF, in reference to a mixer) or baseband (in reference to a modulator) signal can be retained by employing a simple feedback technique known as a series/shunt feedback amplifier, resulting in a significant improvement in the third-order intermodulation (IP_3) and compression point (P_{1dB}) of the mixer.

DIODE RING MIXERS

Diode ring mixers have been widely used since their introduction in the late 1940's, and their nonlinear characteristics were immediately recognized.[3,4] This phenomenon continues to enjoy a very thorough treatment in the professional literature.[5,6,7]

Figure 1 schematically illustrates a form of diode ring mixer commonly referred to as Class I. Here, the four diodes are arranged in a ring, and are switched alternately into on and off states by the application of a local oscillator (LO) signal as shown. Such mixers typically require an LO level of +7dBm (5mW), and subsequent classes require LO levels of +17dBm (50mW) or more, the primary purpose of which is to attain a higher level of IMD performance.

For comparison purposes, a rather common Class I diode ring mixer, the Mini-Circuits SBL-1, was evaluated for intermodulation and compression performance. The SBL-1 enjoys a

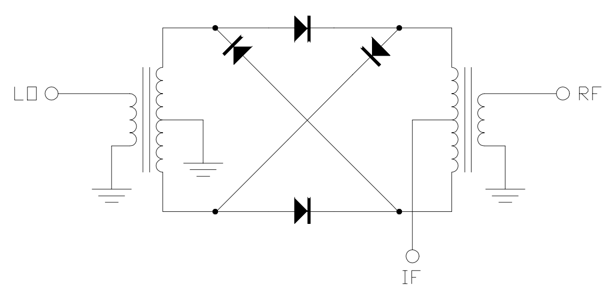

Figure 1 - Typical Class I Diode Ring

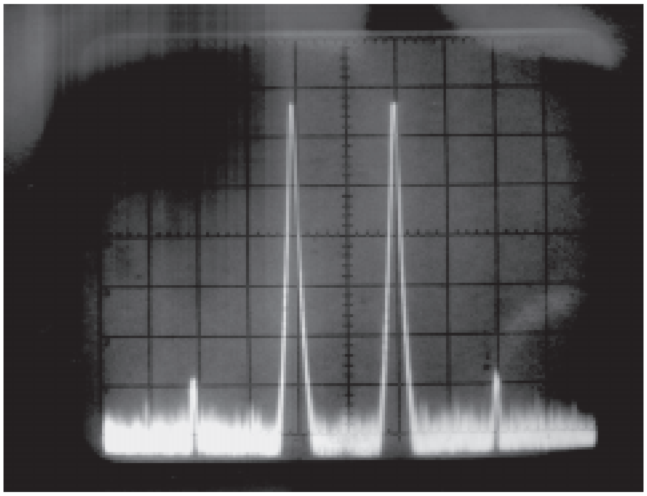

Figure 2 - Intermodulation Distortion
Mini-Circuits SBL-1 Mixer
Ref = 0dBm, 10dB/DIV

Table 1

Input signals	
f_1	500kHz
f_2	510kHz
Power	-9dBm
Local Oscillator	
f_{LO}	10MHz
Power	+7dBm
Output Signal Power	
$f_{LO} + f_1$	-14dBm
$f_{LO} + f_2$	-14dBm
$f_{LO} + 2f_1 - f_2$	-56dBc
$f_{LO} + f_1 - 2f_2$	-56dBc
Gain	-5dB
IIP_3	+19dBm
P_{1dB}	-4.5dBm

wide popularity in radio amateur design, and its commercial counterpart, the SBA-1, is even more widely used, hence its selection for this study.

In the conduction of the testing, the LO frequency is set to 10MHz at the required level of +7dBm. Two IF signals, the first at 500kHz and the second at 510kHz, are applied at the IF port. These signals were chosen as they are well within the capabilities of the device, and will be used in subsequent tests for comparison.

Figure 2 and Table 1 illustrate and tabulate, respectively, the various performance charateristics of the SBL-1. This is fairly typical performance for a Class I diode ring mixer, but as will be shown later, higher IIP_3 and P_{1dB} levels can be attained with a substantially lower LO drive level using an active mixer with a pair of embedded feedback amplifiers.

SWITCHING MOSFET MIXERS

A very worthy variation of the ring mixer uses switching MOSFET devices in place of the diodes, a typical schematic of which is shown in Figure 3. Mixers of this sort often achieve input intercept points (IIP_3) in excess of +40dBm, but do so at the cost of very high LO power levels, usually +17dBm or more, which often makes their implementation in portable battery-powered communications equipment impractical. This, however, is an improvement over Class III diode ring mixers. The professional and hobby literature abound with discussions on the variatious forms of this theme[8, 9, 10, 11, 12, 13, 14], and it would be difficult to give proper attention to this genre without detracting from the purpose of this paper.

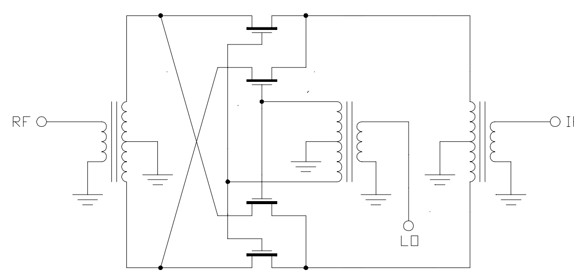

Figure 3 - Ring Mixer with Switching MOSFETs

TRANSITOR TREE MIXERS

Figure 4 schematically illustrates the functional components of a transistor tree mixer. Originally patented in 1966 by Howard Jones as a

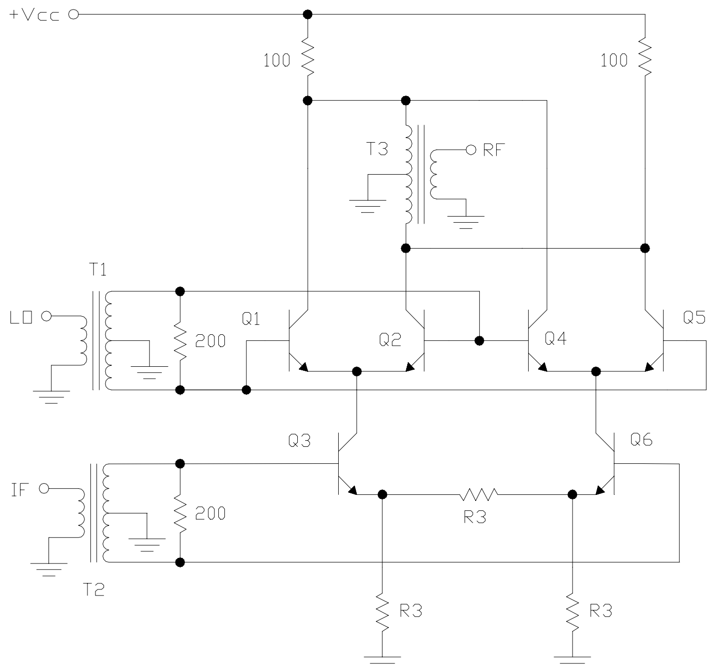

Figure 4 - Transistor Tree (aka Gilbert Cell) Mixer

synchronous demodulator[15], this highly popular active mixer has been more commonly referred to as a Gilbert Cell mixer, following a subsequent patent and it's usage as the core of an analogue multiplier[16]. The tree mixer was derived from earlier synchronous demodulators utilizing vacuum tubes[17]. Here, an input intermediate frequency (IF) signal is applied differentially through transformer T2 across a pair of driver transistors, Q3 and Q6, which generates a differential pair of currents. The resistor network consisting of the three resistors labeled R3 serves to degenerate the emitters of these transistors, which helps stabilize the mixer gain over environmental extremes and also serves to improve the linearity by virtue of being in series with the nonlinear emitter resistances of the driver transistors Q3 and Q6.

These differential currents are then switched alternately by two differential pairs of switching transistors, Q1 and Q2, and Q4 and Q5, respectively, which are driven into alternate on and off states by way of an LO signal applied through transformer T1. By virtue of the summation that takes place through the interconnection of the four collectors of these switching pairs, the LO and IF signals are cancelled at the 100 ohm load resistors, leaving a differential radio frequency (RF) signal across the primary of transformer T3.

For test purposes, a circuit as shown in Figure 4 was constructed using a Harris CA3054 dual differential amplifier array, thus ensuring a good match between all active devices. With a supply voltage of 12 volts, the bias conditions were set to 15mA for the collectors of Q3 and Q6, with the bases of Q3 and Q6 set to 2.1 volts and the bases of the four switching transistors Q1, Q2, Q4, and Q5 set to 4.7 volts, thus ensuring that the transistors Q3 and Q6 will remain in a linear region throughout the range of testing[18].

The resistors R3 are each 100 ohms (a three-resistor array is used). Transformers T1, T2, and T3 are each four turns of trifilar wire through a Fair-Rite 2843-002-402 balun (or binocular) core. With the three transformers having a 1:1:1 turns ratio, the mixer has input and output impedances of 50 ohms at all three terminals.

The same test conditions as used for the diode ring mixer were used here, except that the LO level is now set to 0dBm (1mW). It was determined through testing that the active mixers used in this paper all functioned fully with LO drive levels as low as -6dBm (0.40mW).

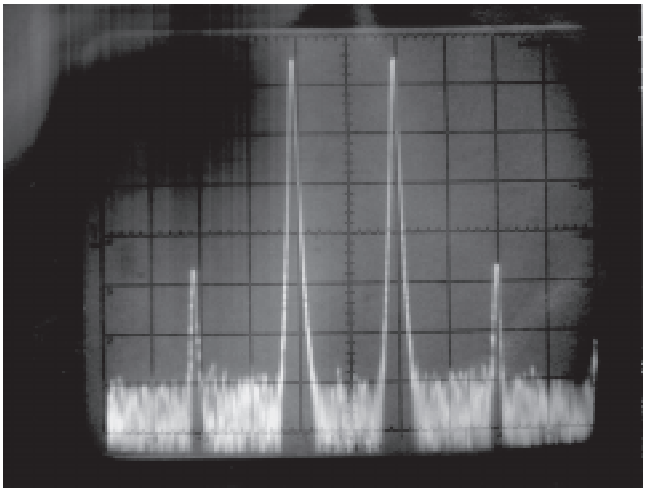

Figure 5 - Intermodulation Distortion
Transistor Tree Mixer
Ref = 0dBm, 10dB/DIV

Table 2

Input signals	
f_1	500kHz
f_2	510kHz
Power	-7dBm
Local Oscillator	
f_{LO}	10MHz
Power	0dBm
Output Signal Power	
$f_{LO} + f_1$	-5.5dBm
$f_{LO} + f_2$	-5.5dBm
$f_{LO} + 2f_1 - f_2$	-42.5dBc
$f_{LO} + f_1 - 2f_2$	-42.5dBc
Gain	-1.5dB
IIP_3	+17.5dBm
P_{1dB}	+4.5dBm

Figure 5 and Table 2 illustrate and tabulate, respectively, the performace of the transistor tree mixer, and the results show that although the 1dB compression point (P_{1dB}) is higher than for the diode ring mixer, the input intercept point (IIP_3) is lower. However, despite the fact that the LO power for the tree mixer is substantially lower than that for the diode ring mixer, the IMD performance is only slightly lower.

It has long been recognized that the most serious limitation in the IMD performance of tree mixers is that of the voltage-to-current conversion of the driver transistors Q3 and Q6.[19, 20] Various methods have been utilized sucessfully to correct this deficiency,[19, 21, 22] but these methods all ignore secondary sources of intermodulation, primarily the h_{fe} nonlinearity of the driver transistors and the nonlinear charcteristics of the four switching transistors. These deficiencies can be overcome by making use of a simple series/shunt feedback amplifier circuit wherein all of the transistors are embedded within the feedback topology.

SERIES/SHUNT FEEDBACK AMPLIFIER

Referring to Figure 5, a series/shunt feedback amplifier is realized by placing a series feedback

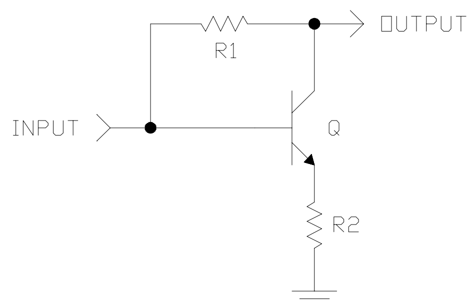

Figure 6 - Series/Shunt Feedback Amplifier

resistor (R2) from the emitter of transistor Q to ground, and a shunt feedback resistor (R1) from the collector to the base. Here, the input and output impedances are determined by the relationship[23, 24]:

$$R_i = R_o = \sqrt{(R1 \times R2)} \qquad (1)$$

and the power gain is:

$$G = \sqrt{[(R1 / R2) - 1]} \qquad (2)$$

This amplifier topology offers a simple means of linearization and is easily implemented in the transistor tree mixer.

LINEARIZED ACTIVE FEEDBACK

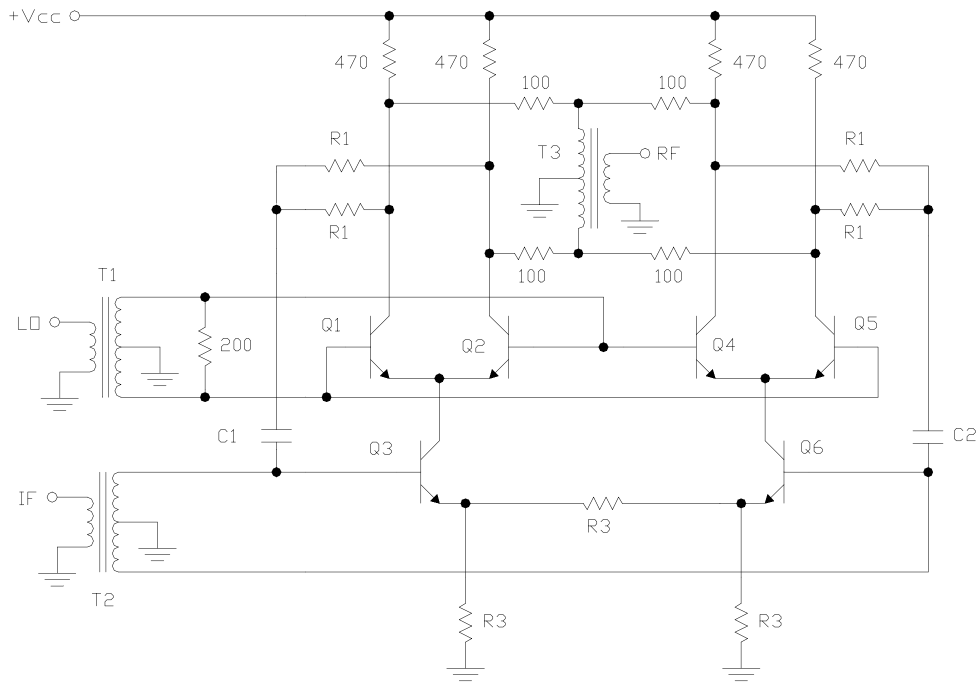

Figure 7 - Linearized Active Mixer Schematic (First Generation)

MIXER (FIRST GENERATION)

Referring to Figure 7, a first series/shunt amplifier is realized by placing a separate shunt feedback resistor from the collectors of the switching transistor pair Q1 and Q2 to the base of the driver transistor Q3. The series feedback resistances are provided by the network of the three R3 resistors. The result here is that the amplified IF signals which are cancelled in the tree mixer are now recovered, while at the same time the RF and LO signals are cancelled at the base of Q3. Thus, the feedback amplifier sees only the input and amplified output IF signals, and since it encompasses all three transistors it can therefore compensate for the distortions caused by the nonlinearities encountered therein.

Similarly, feedback resistors from the collectors of the second switching pair Q4 and Q5 to the base of the second driver transistor Q6, along with the R3 resistor network, completes a second series/shunt feedback amplifier circuit. Note that the value of R3 is three times the value of R2 determined in Equations 1 and 2. The capacitors C1 and C2 provide DC blocking for biasing purposes.

Table 3

Input signals	
f_1	500kHz
f_2	510kHz
Power	-3dBm
Local Oscillator	
f_{LO}	10MHz
Power	0dBm
Output Signal Power	
$f_{LO} + f_1$	-10dBm
$f_{LO} + f_2$	-10dBm
$f_{LO} + 2f_1 - f_2$	-49dBc
$f_{LO} + f_1 - 2f_2$	-49dBc
Gain	-7dB
IIP_3	+21.5dBm
P_{1dB}	+5.5dBm

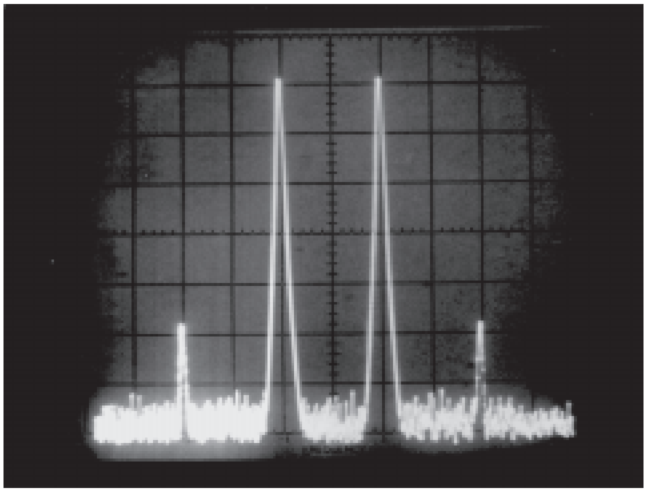

Figure 8 - Intermodulation Distortion
Linearized Active Mixer (First Generation)
Ref = 0dBm, 10dB/DIV

The network of four 100 ohm resistors from the four switching transistor collectors to the output transformer serves to cancel the LO and IF signals at the output transformer primary, while at the same time providing a differential RF output signal.

A test circuit was constructed using the same components and bias conditions as for the earlier tree mixer. The four resistors R1 were set to 330 ohms, giving both amplifiers an input and output impedance of 100 ohms and an IF signal gain of 6.7dB.

The test results shown in Figure 8 and Table 3 indicate that the feedback mixer circuit of Figure 7 has measurably better IMD performance than that of the comparable tree mixer of Figure 4 and also outperforms the Mini-Circuits SBL-1 diode ring mixer while requiring substantially less LO power. The compression point (P_{1dB}) suffers slightly as a result of incomplete cancellation of the LO signal at the collectors of the four switching transistors, which leads to early saturation of these devices.

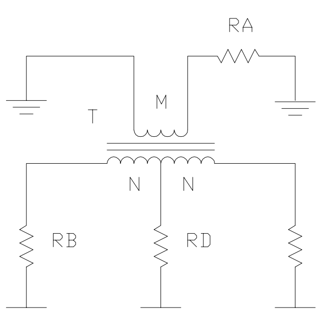

Figure 9 - Hybrid Transformer

The four 100 ohm resistors in the output network are at fault here, as they provide a lossy path between the opposite collectors which, in a tree mixer, provide for this cancellation. They also result in an unnecessary loss of 6dB in output signal power. A remedy of this shortcoming was found in the form of a signal combining device known as a hybrid transformer.

HYBRID TRANSFORMER COMBINERS

Hybrid transformers[25, 26, 27] (also referred to as bridge transformers or balanced transformers) are magnetic devices commonly found in telephone repeater amplifiers, but which, with proper materials, are readily applied to higher frequency circuitry. Refer-ring to Figure 9, the hybrid transformer discriminates between common-mode and odd-mode signals. Common-mode signals across the centre-tapped primary are isolated from the secondary and appear at the centre tap. Conversely, odd-mode signals are isolated from the centre tap and appear at the secondary winding. The primary and secondary have turns of 2N and M, respectively. The four impedances are determined by:

$$RD = (N / M)^2 \times RA \qquad (3)$$

$$RB = RC = 2 \times RD \qquad (4)$$

In the previous realization, the resistors used in the output signal combiner resulted in a 6dB loss in gain. Use of the hybrid combiner negates this loss, and thus the use of the term "lossless" when referring to this topology.

LINEARIZED ACTIVE MIXER WITH LOSSLESS HYBRID COMBINERS (SECOND GENERATION)

Referring now to Fig. 10, a functional schematic of a linearized double-balanced active mixer with lossless hybrid combiners is shown. The

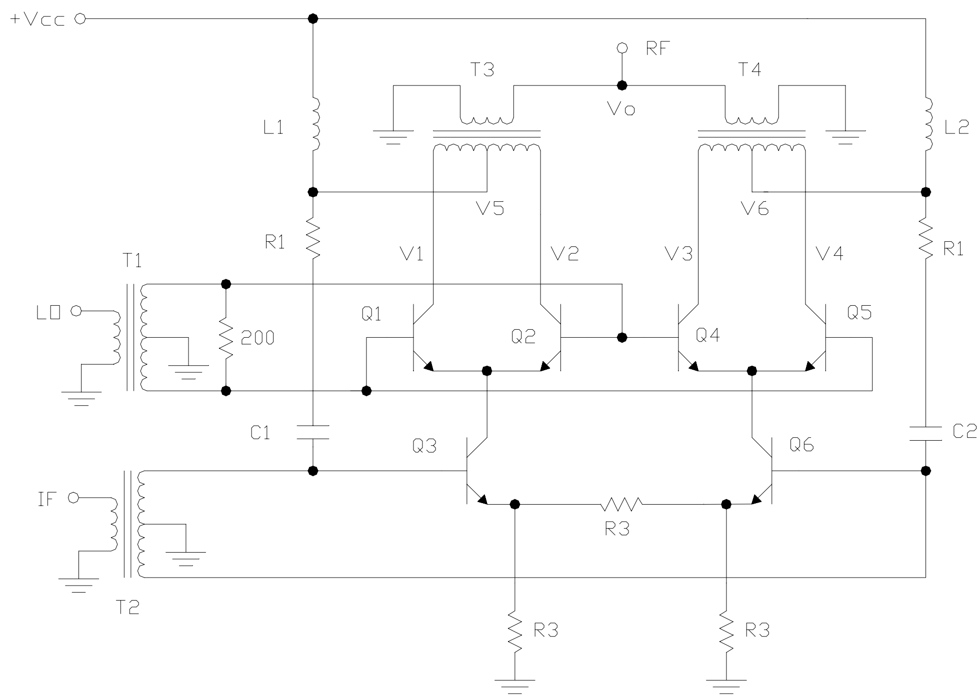

Figure 10 - Linearized Active Mixer Schematic (Second Generation)

circuit consists of two balanced active mixers, and the discussion will be limited to the side to the left, there being no unique discussion for the right side.

To begin, the mixer as a whole sees an RF load impedance of R_L, and each side of the mixer sees a load impedance of $2R_L$. With a hybrid transfomer having a turns ratio of 1:1:1, the centre tap of the primary will also have an impedance $2R_L$, while the end terminals of the primary will have an impedance of $4R_L$. The switching action of transistors Q1 and Q2 will modulate the collector current of transistor Q3, creating a differential signal across the primary winding of T3. Since the driver transistor Q3 sees a constant collector load, it's overall load impedance is equal to the parallel combination of the collector loads of Q1 and Q2, which is the the same as the impedance of the primary centre tap, $2R_L$. Thus, the series/shunt feedback amplifier is again realized.

The four collector voltages for transistors Q1, Q2, Q4, and Q5 are, respectively:

$$V1 = -A \times A_V \times [1/2 \cos(\omega_L - \omega_S) + 1/2 \cos(\omega_L + \omega_S)] - I \times R_L \cos \omega_L \qquad (5)$$

$$V2 = A \times A_V \times [1/2 \cos(\omega_L - \omega_S) + 1/2 \cos(\omega_L + \omega_S)] + I \times R_L \cos \omega_L \qquad (6)$$

$$V3 = A \times A_V \times [1/2 \cos(\omega_L - \omega_S) + 1/2 \cos(\omega_L + \omega_S)] + I \times R_L \cos \omega_L \qquad (7)$$

$$V4 = -A \times A_V \times [1/2 \cos(\omega_L - \omega_S) + 1/2 \cos(\omega_L + \omega_S)] - I \times R_L \cos \omega_L \qquad (8)$$

where A is the IF input signal level, A_V is the amplifier gain from (2), ω_L is the input local oscillator (LO) frequency, ω_S is the IF input signal frequency, and I is the quiescent collector bias current of Q3. The third term in equations (5) and (6) represents a differential carrier signal across the primary of T3. There is an equal, but opposite, signal across T4 (equations 7 and 8), effectively cancelling the two LO signals, and the balance of these two signals will determine the re-sulting LO/RF leakage. Under ideal (ie - lossless) con-ditions, the four voltages above now become:

$$V1 = -A \times A_V \times [1/2 \cos(\omega_L - \omega_S) + + 1/2 \cos(\omega_L + \omega_S)] \quad (9)$$

$$V2 = A \times A_V \times [1/2 \cos(\omega_L - \omega_S) + + 1/2 \cos(\omega_L + \omega_S)] \quad (10)$$

$$V3 = A \times A_V \times [1/2 \cos(\omega_L - \omega_S) + + 1/2 \cos(\omega_L + \omega_S)] \quad (11)$$

$$V4 = -A \times A_V \times [1/2 \cos(\omega_L - \omega_S) + + 1/2 \cos(\omega_L + \omega_S)] \quad (12)$$

The recovered IF signals V5 and V6 at the centre taps of hybrid transformers T3 and T4 are:

$$V5 = -A \times A_V \times \cos \omega_S \quad (13)$$

$$V6 = A \times A_V \times \cos \omega_S \quad (14)$$

and the RF output voltage is:

$$V_O = (M/N) \times A \times A_V \times [\cos(\omega_L - \omega_S) + + \cos(\omega_L + \omega_S)] \quad (15)$$

which, if M = N, becomes

$$V_O = A \times A_V \times [\cos(\omega_L - \omega_S) + + \cos(\omega_L + \omega_S)]$$

(16)

A test circuit was constructed, again using the same devices and bias conditions as previous. Referring to Figure10, the two hybrid transformers T3 and T4 are of the same construction as the input transformers T1 and T2, having turns ratios of 1:1:1 and consisting of four turns of trifilar wire on a Fair-Rite 2843-002-402 two-hole balun core. Thus, each side of the mixer sees a source and load impedance of 100 ohms, and by virtue of T4 and the paralleled secondaries of T3 and T4, the input and output impedance of the mixer is 50 ohms.

Testing was conducted using the same signals as before. Figure 11 and Table 4 illustrates and tabu-lates, respectively,the performance of this mixer. With the third-order intemodulation prod-ucts at -53dBc, the resulting input intercept point is a respectable +29.5dBm. Also, the compression point has now risen to +10.5dBm The inclusion of the hybrid transformers has here produced an active mixer with IMD performance rivaling that of Class

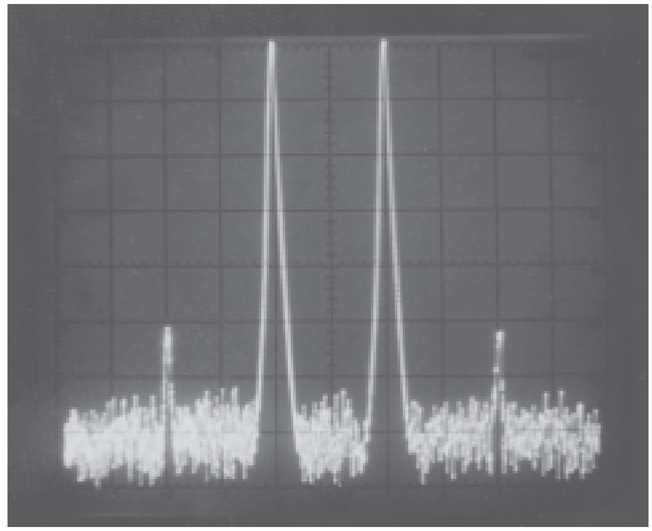

Figure 11 - Intermodulation Distortion Linearized Active Mixer (Second Generation)

Table 4

Input signals	
f_1	500kHz
f_2	510kHz
Power	+3dBm
Local Oscillator	
f_{LO}	10MHz
Power	0dBm
Output Signal Power	
$f_{LO} + f_1$	0dBm
$f_{LO} + f_2$	0dBm
$f_{LO} + 2f_1 - f_2$	-53dBc
$f_{LO} + f_1 - 2f_2$	-53dBc
Gain	-3dB
IIP_3	+29.5dBm
P_{1dB}	+10.5dBm

	Mini-Circuits SBL-1	Transistor Tree Mixer	Linearized Feedback Mixer (1st Generation)	Linearized Feedback Mixer (2nd Generation)
Without IF Filter				
P_{1dB}	–4.5dBm	+4.5dBm	+5.5dBm	+10.5dBm
IIP_3	+19dBm	+17.5dBm	+21.5dBm	+29.5dBm
With IF Filter				
P_{1dB}	–7.5dBm	+4.5dBm	+5.5dBm	+10.5dBm
IIP_3	+7.5dBm	+16.5dBm	+20.75dBm	+28.5dBm

Table 5 - Narrow-Band IF Load Performance

III diode ring mixers, but with a fraction of the LO drive power.

IF LOAD SENSITIVITY

One aspect of mixer performance that bears close scrutiny is the susceptibility to narrow-band IF loads. In the design of radio transmitter and receiver systems, the rejection of images and spurious responses from frequency conversion requires that the output of the mixer be filtered, more often than not by way of a bandpass filter. Diode ring and switching MOSFET mixers are notoriously sensitive to such loading impedances, and the worst case, at least for active mixers, is that in which the unwanted signal voltage is terminated in a high impedance.

To this effect, a coupled resonator bandpass filter was constructed with a centre frequency of 10.7MHz and a passband bandwidth of 500kHz, the schematic of which is shown in Figure 12. Insertion loss was measured at 5.5dB, which is taken into account in the measured data.

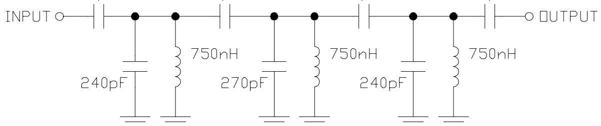

Figure 12 - Bandpass FIlter Used for IF Load Tests

The test results, shown in Table 5, show that the SBL-1 diode ring mixer is indeed seisitive to the narrow-band IF termination, the IIP_3 decreasing by 11.5dB and the compression point by 3dB. The active mixers, without exception, show a substantially re-duced sensitivity, with the compression points remaining unaffected and the IIP_3 cdecreasing 1dB or less in all three cases.

This is not to be unexpected: In the case of the diode ring mixer, the unterminated signal energy is reflected back to the diodes where it can further interact with the nonlinearity of the diode junctions. On the other hand, the signal energy reflected back to the active mixers is terminated in the switching transistor load resistances, the nonlinear base-emitter junctions being isolated by virtue of the reverse transfer coefficients of the transistors.

CONCLUSION

An active mixer of improved performance over a previously disclosed design has been demonstrated and shown to be have charactereistics that are desirable in the design of high performance RF receiver and transmitter systems. Further improvements are envisioned, including the usage of alternative feedback topologies that will address the noise figure characteristics, rendering a mixer of very high dynamic range without the need for exorbitant LO power levels.

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