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Title: Modern Receiver Mixers For High Dynamic Range
Author: Doug DeMaw, W1FB

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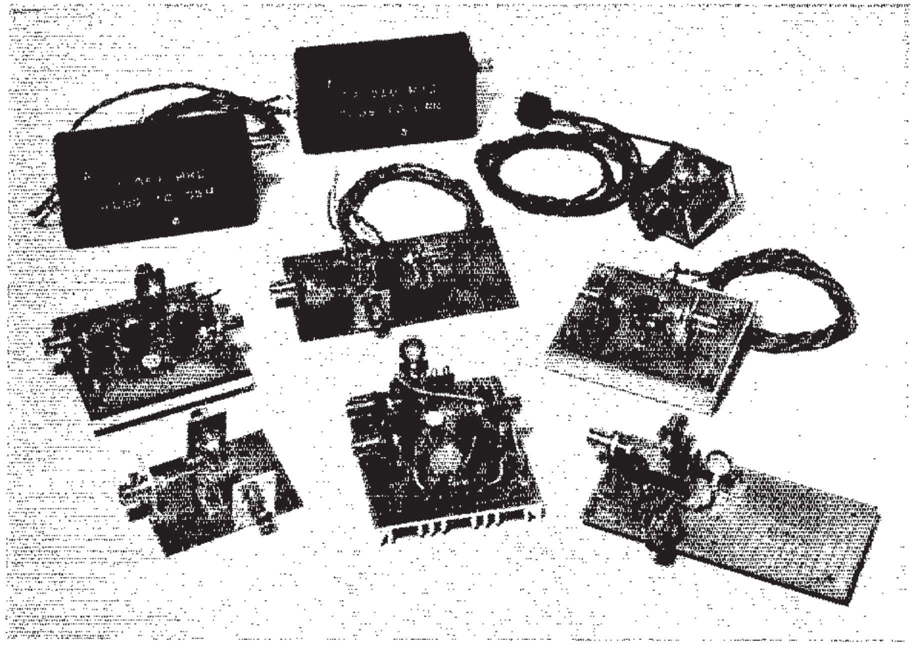
Modern Receiver Mixers for High Dynamic Range



The VMP4 VMOS offers a new approach to mixer design. For quality communications equipment it provides improved port-to-port isolation and relatively high immunity to strong in-band and out-of-band signals.

By Doug DeMaw,* W1FB and George Collins,** ADØW

Designers and users of modern communications receivers and transmitters are necessarily interested in high dynamic range and port-to-port isolation in the mixer stages of the equipment. A quality communications receiver for medium- or high-frequency band use will exhibit high dynamic range in order to provide relative immunity to strong in-band or out-of-band signals. Furthermore, the system should be relatively free of spurious responses that cause "birdies" across the receiver tuning range.

Acknowledging the importance of gain distribution and noise figure in the early stages of a receiver, we concentrate, therefore, on the mixer performance. In a typical quality design of the day we try to ensure a "crunch-proof" status for the rf amplifier, mixer and post-mixer amplifier. For the most part, this requires that each of those stages be capable of handling a substantial amount of signal power without gain compression or undue IMD products being generated. It is not unusual to find a VMOS power FET or a large CATV (cable television) type of bipolar transistor being used as an rf amplifier ahead of the mixer. A VMP4 VMOS device or a 2N5109 bipolar transistor can be used to obtain high dynamic performance in an rf amplifier. The same or similar devices are often used as post-mixer broadband amplifiers in high-performance communications receivers.[1,2]

Our objective is to select a mixer that has sufficient port-to-port isolation to minimize the effects of LO energy appearing in the mixer output. Similarly, the signal energy should be well suppressed at the remaining ports of the mixer. Furthermore, if the mixer requires a high level of LO power to provide optimum performance, difficulty may be encountered in keeping the LO energy isolated from the other circuits in the receiver. Our choice, therefore, must be one that involves a minimum amount of trade-offs while ensuring good mixer performance.[3]

Mixer Options

The choice between passive and active mixers in a given design should be based on performance objectives, with consideration of the circuitry that precedes and follows the mixer. A singly or doubly balanced mixer is preferable to a single-ended mixer in the interest of isolation between the ports. The doubly balanced version will afford the best performance in that respect.

The active mixer will yield conversion gains of less than unity to as great as 20 dB, depending upon how it is used. Perhaps the least acceptable of the better mixer options is a pair of small-signal, dual-gate MOSFETs of the 40673 family. Many communications receivers use such devices in a broadband, singly balanced arrangement. Although this may be cost-effective to the manufacturer, high dynamic range will be hard to achieve without a sacrifice in noise figure at the higher frequencies. If an rf amplifier is used to improve the noise performance, care must be taken to keep the gain only

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[1]References appear on page 23.

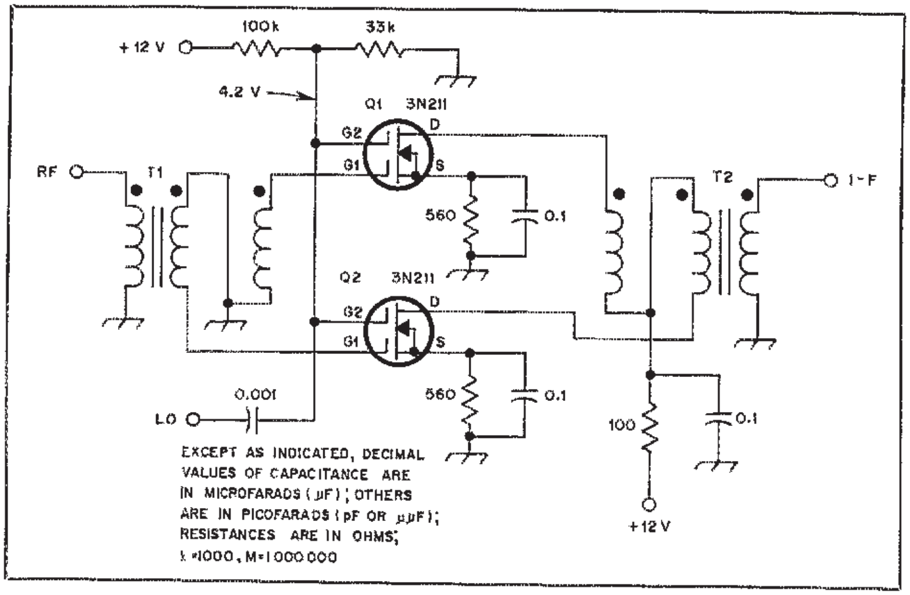


Fig. 1 — Reference mixer that uses small-signal dual-gate MOSFETs in a broadband, singly balanced setup. Conversion gain is −5 dB because of the low terminal impedances and broadbanding. With an LO injection of 8 volts pk-pk and an input signal level of −10 dBm PEP, the third-order output intercept is +17 dBm. Gate bias and LO injection has been optimized. Narrowbanding and careful impedance matching would yield conversion gains up to +15 dB.

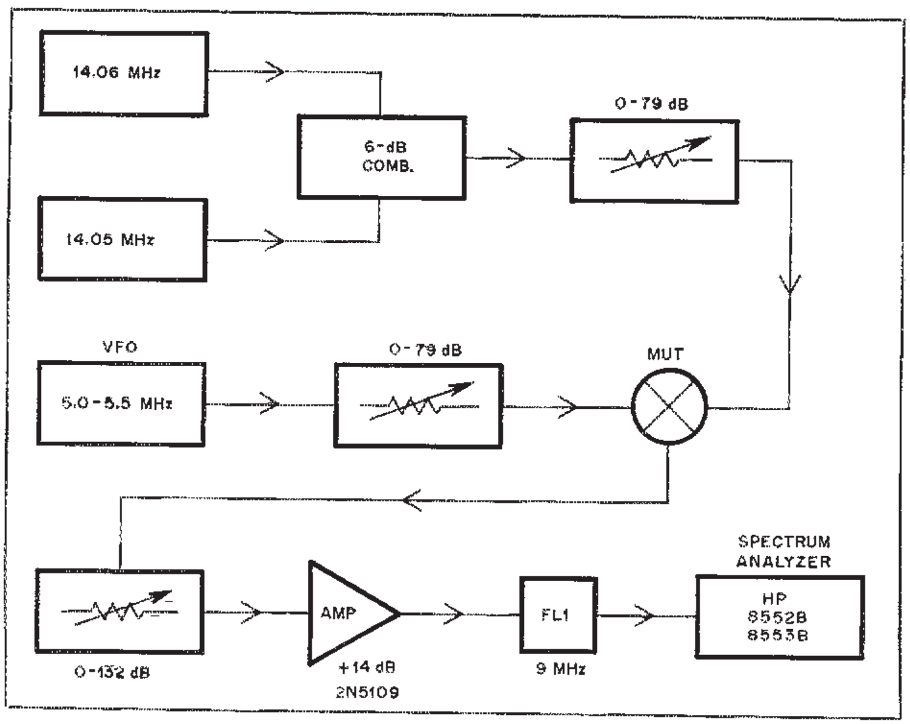


Fig. 2 — Block diagram of the test fixture used in evaluating the mixers treated in this article.

high enough to ensure the desired receiver noise figure. Too much gain will cause the mixer to collapse in the presence of strong signals. Most receivers that use small-signal FETs in a singly balanced scheme (Fig. 1), and with a MOSFET rf amplifier ahead of the mixer, exhibit an IMD characteristic of 80 to 85 dB if the design has been done with care. A gain compression of 1 dB occurs between 115 and 125 dB on the average, referenced to the receiver MDS (minimum discernible signal). A receiver with these characteristics might be entirely acceptable in some signal environments. But much greater dynamic range is necessary in high signal-density locations, such as shipboard and in large communications centers where transmitters are operating on several frequencies simultaneously.

Active mixers can be valuable in terms of conversion gain, with narrow-band types yielding the higher gain figures. The usual trade-off between bandwidth and gain must be accepted when using broadband mixers. The amount of conversion gain desired will depend on the filter losses before and after the mixer, and the available overall gain after the mixer.

Passive mixers of the diode-ring, doubly balanced variety are capable of excellent dynamic range and port-to-port signal isolation. The shortfall is, of course, fairly high LO power requirements (+7 to 15 dBm) and a conversion loss on the order of 8 dB. It is almost mandatory to employ an rf amplifier ahead of the diode-ring mixer to provide a low noise figure. A diplexer may be used after the mixer to ensure a 50-ohm termination at all frequencies, thereby aiding the IMD characteristic.[4] If the i-f filter has a high insertion loss (10 dB for most mechanical filters), a post-mixer, large-signal amplifier is worthwhile. It can be terminated by a 50-ohm pad of specified attenuation to ensure a constant load and to protect the i-f filter from damage when very high signal levels are present at the receiver input.

It is apparent from the foregoing discussion that a lot of decision making is necessary when deciding what mixer to use. Whatever the choice is, high dynamic range should be the criterion. This can be achieved with passive or active mixers. The remainder of this article addresses various mixers and their performance characteristics. The laboratory test procedure used by the authors is also discussed.

Mixer Evaluation Method

Two-tone tests of the various mixers were performed with a signal separation of 10 kHz at 14.050 and 14.060 MHz. A +8 dBm output level was available from each spectrally clean, crystal-controlled generator. A 6-dB combiner followed the two generators. Output from the combiner was routed through a Tektronix 2701 attenuator (0 to 79 dB), then to the mixer under test (MUT).

LO power was generated by a Trio-Kenwood 5-MHz VFO, to which filtering and additional amplification was added. The LO source delivered +16 dBm. A second Tektronix 2701 attenuator was connected between the LO source and the MUT LO port.

I-f output from the mixers was routed through an HP 355C/D attenuator (0 to 132 dB) to a broadband class A 2N5109 +14-dB amplifier which had a +40 dBm output intercept. A 2.4-kHz bandwidth KVG 4-pole crystal-lattice filter with an IL (insertion loss) of −5 dB followed the post-mixer amplifier. Output from the filter was supplied to an HP 8553/8552B spectrum analyzer through a 500- to 50-ohm matching transformer. Other broadband transformers were used to provide a proper interface between the test modules and MUTs. Those transformers are not shown in the test-setup block diagram of Fig. 2. A 7-pole, T-section Chebychev low-pass filter was used at the output of the 2N3866 broadband post-LO amplifier to ensure that all LO harmonics were 70 dB or greater below peak LO output.

Plessey SL6440C IC Mixer

A recent product to the IC market is the Plessey SL6440C programmable high-level mixer. It is advertised as having a +30 dBm output intercept point and a +15 dBm compression point (1 dB). The internal circuit of the IC had not been revealed at the time this article was written, but it is presumed that the inner workings are not too unlike those of the Motorola MC1496G, with the exception of greater dissipation capability for the SL6440C. The manufacturer rates the mixer as having a −1 dB (typical) conversion gain when the IC is terminated in 50 ohms. In our tests a 200-ohm termination was used at the input and output of the IC, yielding a conversion gain on the order of +8 dB maximum.

Fig. 3 contains the circuit of the SL6440C as it was configured for laboratory analysis. R1 was used to adjust the quiescent current of the mixer. Table 1 shows the test results at various LO and signal-input levels. The spectral displays of Fig. 4 show the LO and LO harmonics to i-f-port isolation (A). With 0 dBm of LO input power the isolation was 29 dB. The 2f LO isolation was measured at −72 dB.

Spectral photograph B of Fig. 4 shows the rf port to i-f port isolation as being 48 dB when the LO level was 0 dBm and the rf-input level was −10 dBm, PEP. Photograph C is the two-tone output of the mixer. Table 2 contains data on conversion gain, intercept numbers and port isolation with various I_p amounts. These data were compiled with an LO injection of 0 dBm and an rf input of −5 dBm.

One might conclude from the foregoing test results that the Plessey SL6440C is indeed a worthy device which is capable of providing high dynamic range without conversion loss. The I_p values used in these tests were the maximum safe levels

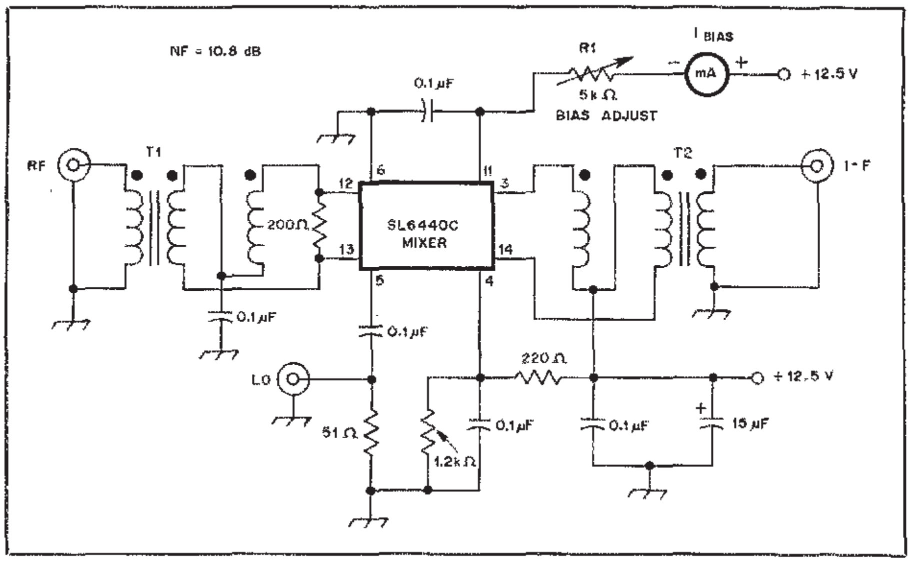


Fig. 3 — Test circuit for the Plessey SL6440C mixer IC.

Table 1

Test Results at Various LO and Signal-Input Levels

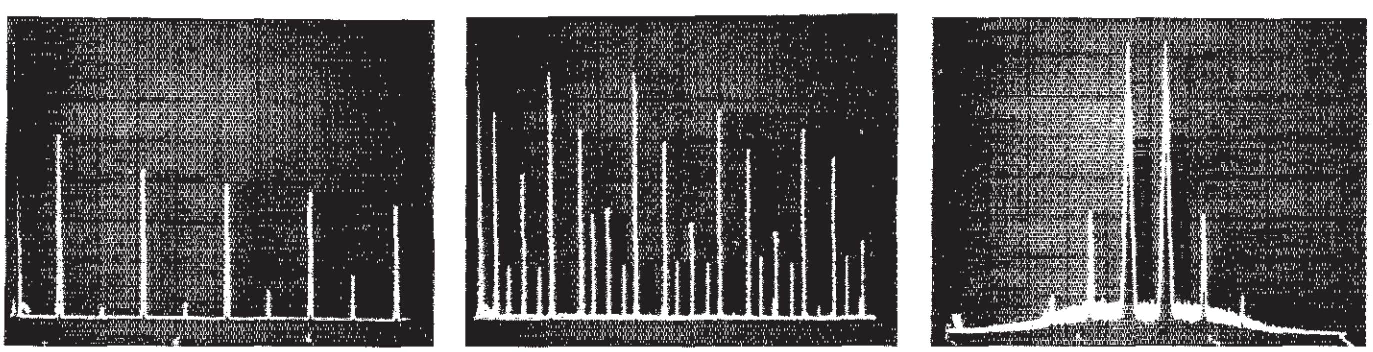
LO Level (dBm)	Input Level (PEP, dBm)	Third-Order Output Intercept (PEP, dBm)	Conversion Gain (dB)
0	+3	24	7
0	0	29	8
0	−5	31	8
0	−10	31	8
0	−15	29	8
−3	+3	25	7
−3	+1	28	8
−3	0	29	8
−3	−5	30	8
−3	−10	31	8
−3	−15	30	8
−10	+2	23	7
−10	0	29	8
−10	−5	31	8
−10	−10	31	8
−10	−15	29	8
−15	+2	25	7
−15	0	29	8
−15	−5	31	8
−15	−10	31	8
−15	−10	31	8
−15	−15	29	8

Table 2

Data on Conversion Gain, Intercept Numbers and Port Isolation

I_p Current (Pin 11, mA)	Conversion Gain (dB)	Third Order Output Intercept (dBm)	LO-to-RF Isolation (dB)
5.0	5.5	18	27
5.5	6.0	19	27
6.0	6.5	20	27
6.5	7.0	21	27
7.0	7.0	22	27
7.5	7.0	23	27
8.0	7.0	24	27
8.5	7.0	25	27
9.0	7.5	26	27
9.5	7.5	26	27
10.0	7.5	27	27
10.5	7.5	28	27
11.0	7.5	28	27
11.5	8.0	29	27
12.0	8.0	29	28
12.5	8.0	30	28
13.0	8.0	31	28
13.5	8.0	31	28

LO input = 0 dBm. Rf input = dBm



(A) (B) (C)

Fig. 4 — Spectograph A shows the LO suppression at the i-f port of the SL6440C with no rf signal applied. LO injection is 0 dBm. Center frequency is 25 MHz, bandwidth is 100 kHz, vertical scale is 10 dB/div. and horizontal scale is 5 MHz/div. Display B shows the output spectrum with 0 dBm of LO power and −10 dBm of rf signal applied to the mixer. Analyzer bandwidth in this example is 30 kHz. Two-tone output is displayed at C with the vertical scale being 10 dB/div. and the horizontal scale at 10 kHz/div. Center frequency is 9 MHz and bandwidth is 0.3 kHz.

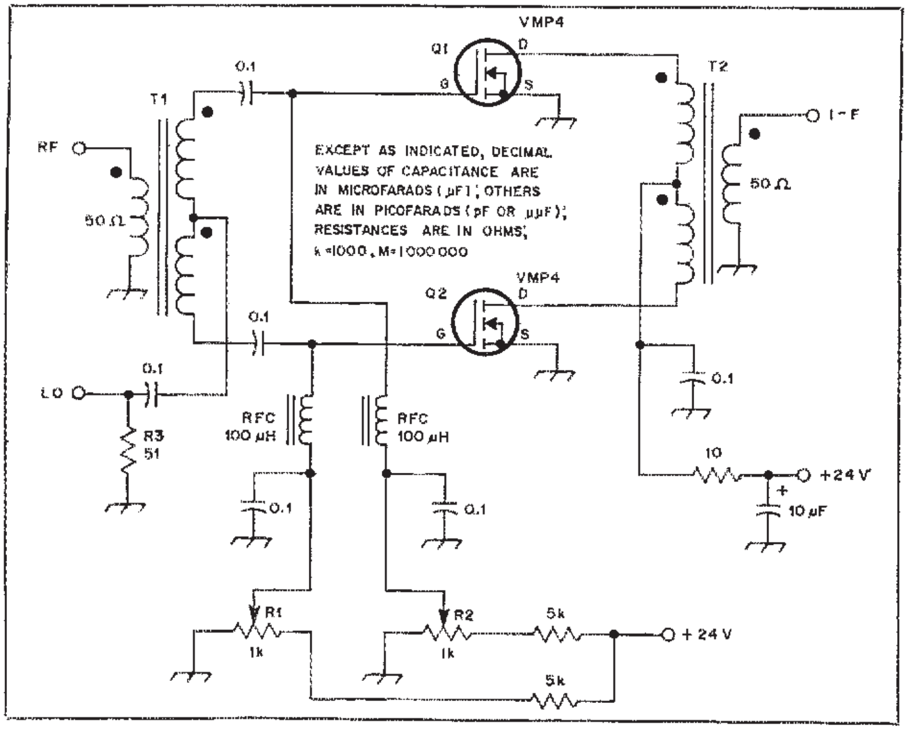


Fig. 5 — Circuit for the VMOS power FETs in a singly balanced arrangement. Split dc feed to the gates was used to provide dynamic balance.

Table 3

Results with Various Levels of Gate Voltage and Drain Currents

LO (dBm)	Rf Input[1] (dBm)	Gain (dB)	I_D[2] (mA)	Gate[3] Volts	Third O.I. (dBm)
+16	+8	15	75	1.0	42.5
+16	+5	15	63	1.0	43.5
+16	+2	15	56	1.0	44.0
+16	−1	15	54	1.0	44.0
+16	−4	15	52	1.0	45.0
+16	−7	15	52	1.0	*
+16	+8	16	115	1.5	42.0
+16	+5	16	105	1.5	44.0
+16	+2	16	100	1.5	45.0
+16	−1	16	97	1.5	45.0
+16	−4	16	96	1.5	44.5
+16	−7	16	96	1.5	*
+16	+8	16	180	2.0	39.0
+16	+5	17	170	2.0	42.5
+16	+2	17	165	2.0	43.0
+16	−1	18	160	2.0	44.5
+16	−4	18	160	2.0	43.0
+16	−7	18	160	2.0	43.0

*IMD products below measurement system noise floor.
[1]All signal levels referenced to PEP.
[2]Total current.
[3]Both gates at same voltage.

of dc current without heat sinking the IC. A package limitation of 1.2 watts is specified for temperatures up to 25° C, with derating set at 8 mW/° C above 25° C. Maximum program current is 50 mA.

Singly Balanced VMOS Power FET Mixer

The VMOS power FET has characteristics that suggest its ability to perform well in a high-level balanced mixer circuit. For this reason it was included in the mixer evaluation program to determine how it would compare to other high-level mixers. A pair of VMP4 vhf devices was selected for testing in a singly balanced mixer. Other VMOS devices, such as the VN66AK, should offer nearly comparable hf-band performance at lower cost. The VMP4s were chosen mainly because they could be adapted easily to heat sinking, owing to the strip line package format.

Fig. 5 contains the test circuit used by the authors. The rf and LO signals were applied to the gates of the FETs, permitting the sources to be grounded. Earlier tests of the same FETs in a circuit recommended by a manufacturer (rf signal injected on the sources and LO injection on the gates) yielded substantially degraded mixer performance. Instability was also manifest when the forward gate voltage was increased beyond 1.9. Stability could not be obtained without excessive resistive loading of the broadband transformers, so the circuit was abandoned in favor of the one in Fig. 5. Bias controls R1 and R2 were included to help establish dynamic balance of Q1 and Q2. R3 was added to establish a known impedance at the LO injection point during laboratory analysis. Without the resistor, the port impedance is in excess of 500 ohms.

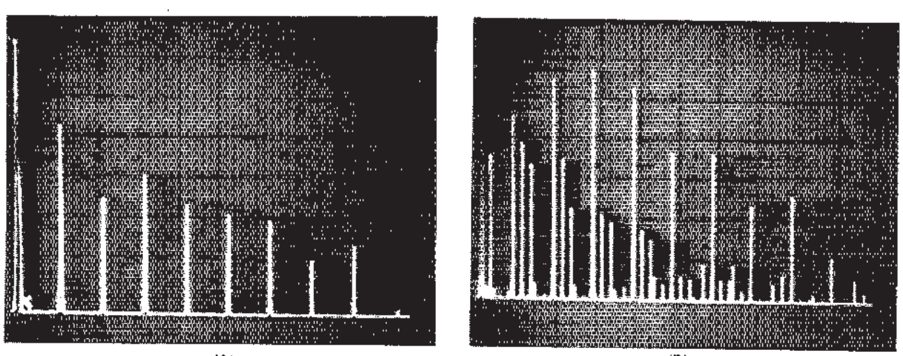
Table 3 shows the results obtained with various levels of gate voltage, rf-signal input and quiescent drain currents. LO injection was maintained at +16 dBm. It can be seen that a variety of operating conditions yielded good output intercepts. The resultant conversion gain is somewhat higher than is desired for most receiver applications. If this circuit is used it will probably require inclusion of an attenuator pad after the mixer to tailor the effective gain to a suitable level for the stages that follow the mixer.

As one would suspect, port isolation follows the format that is common to singly balanced mixers. With the circuit of Fig. 5 the isolation was 38 dB when R1 and R2 were adjusted for best suppression of the output responses. This condition was realized when one gate had 2 volts and the other had 1.85 volts. Fig. 6 shows the spectral output of the mixer under a balanced condition. Photograph A shows the LO isolation and photograph B illustrates the LO and rf isolation.

The authors have concluded that VMOS power-FET mixers are worth considering when high dynamic range is desired (without concern for the high values of quiescent drain current in a 24- to 28-volt dc type of system). It follows that a quad of power FETs in a doubly balanced mixer circuit would offer improved performance over that provided by the mixer in Fig. 5.

High-Level Diode-Ring DBM

If the designer is willing to accept a trade-off between dynamic range and conversion gain, the doubly balanced diode-ring mixer is worthy of consideration. Our



(A) (B)

Fig. 6 — Spectral display A shows the LO/i-f port isolation for the VMP4 balanced mixer with +16 dBm of LO injection and no rf signal applied. Spectrograph B reveals the rf/i-f isolation with +16 dBm of LO power and 0 dBm of rf signal input.

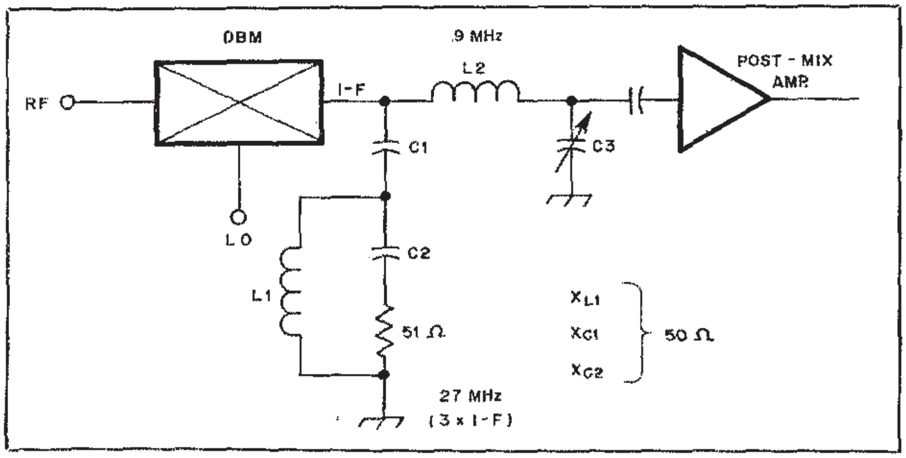


Fig. 7 — Method for adding a diplexer to the output of a diode-ring mixer to enhance the IMD performance. The high-pass network is terminated in 51 ohms and is designed for 3 × i-f. An L network provides an impedance match between the mixer output (50 ohms) and the input of the post-mixer amplifier.

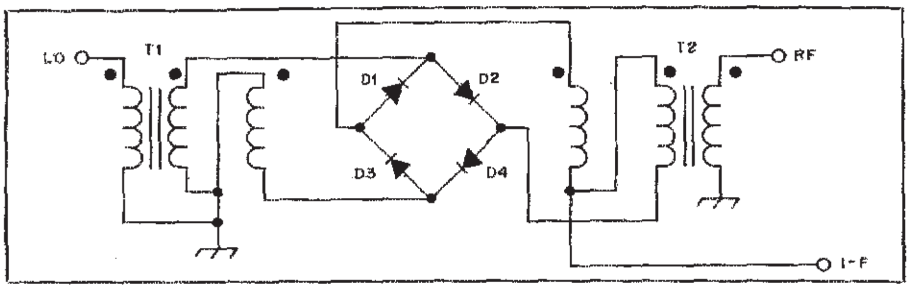


Fig. 8 — Diagram of the SRA-1H diode-ring, high-level mixer used in the performance tests.

Table 4

Test Results for the SRA-1H Diode-Ring High-Level Mixer

LO (dBm)	RF (dBm, PEP)	I-f (dBm, PEP)	Gain (dB)	Third-Order O.I. (dBm, PEP)
+9	+8	0	−8	+19.0
+9	+5	−2	−7	+22.0
+9	+2	−5	−7	+23.5
+9	−1	−8	−7	+24.5
+9	−4	−11	−7	+26.5
+9	−7	−14	−7	*
+12	+8	+1	−7	+25.0
+12	+5	−2	−7	+26.0
+12	+2	−5	−7	+27.0
+12	−1	−8	−7	+27.0
+12	−4	−11	−7	*
+12	−7	−14	−7	*
+15	+8	+1	−7	+30.0
+15	+5	−2	−7	+30.0
+15	+2	−4	−6	+31.0
+15	−1	−7	−6	+31.0
+15	−4	−10	−6	*
+15	−7	−13	−6	*
+17	+8	+1	−7	+33.0
+17	+5	−1	−7	+33.5
+17	+2	−4	−6	+33.0
+17	−1	−7	−6	+31.0
+17	−4	−10	−6	*
+17	−7	−13	−6	*

*Measurement limited by post mixer/filter IMD.

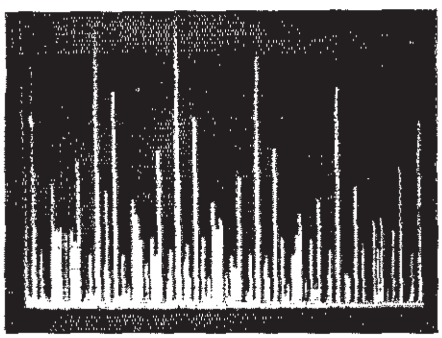


Fig. 9 — Output display of the SRA-1H high-level, diode-ring mixer with an LO power of +17 dBm and an rf signal input of +8 dBm. The large number of spurs emphasizes the importance of filtering at the mixer output.

tests included an analysis of the Mini-Circuits Lab SRA-1H DBM module. The test setup was essentially the same as for the previous mixers treated in this article. Tests were conducted with and without a diplexer connected to the mixer output. The results were essentially identical, since the test-setup terminations provided the desired 50-ohm port characteristic. In an actual receiver where absolute source-impedance levels are not always known, a diplexer of the type shown in Fig. 7 can be beneficial in providing the mixer with a 50-ohm termination at all frequencies. The high-pass branch of the diplexer is resonant at approximately three times the i-f. Improvements of 2 to 3 dB in mixer IMD are not uncommon when a diplexer is added to a ring mixer.

Test results for the SRA-1H (Fig. 8) are listed in Table 4. The conversion-gain spread follows the predicted amount, ranging from −6 to −9 dB over an LO injection excursion of +6 to +17 dBm. A spectral display of the mixer output is shown in Fig. 9. Owing to the LO power needed for this mixer it became necessary to make one change in the test setup used for the other mixers discussed here: The 2N3866 post LO amplifier was followed by an MRF-511 CATV transistor to elevate the available LO power to +27 dBm. The second harmonic from the LO source was measured at greater than 50 dB below the peak power of the fundamental.

Summary

The implications of the test results in this article are that large-signal devices provide high dynamic-range numbers when careful attention is given to biasing and LO levels. Certainly, a receiver is only as good as its mixer in terms of large-signal accommodation. Schottky ring mixers still offer a good compromise between dynamic range and moderate LO injection power. The penalty is in conversion loss, but a major advantage is seen in the passive feature of the diode-ring mixer, since the device does not impose dc current drain on the power supply.

The Plessey SL6440C high-level mixer IC offers the advantage of having excellent dynamic range, conversion gain and moderate dc current requirements. The LO-injection level is significantly lower than that required for a diode-ring mixer, and the package format lends itself well to the design of miniature equipment.

VMOS power FETs open the door to very high dynamic-range numbers at the cost of bulk and high dc-current requirements. LO-injection requirements are fairly high, and heat sinking of the active devices is necessary. Owing to the relatively high amplitude of the mixer-output spurs, a doubly balanced VMOS mixer would be preferable to a singly balanced version. The VMOS balanced mixer may not, in many instances, be cost-effective unless low-priced VMOS FETs are used in preference to the VMP4s specified in this article. The latter are in the $20 price class when purchased in single-lot quantity.

Indications are that small-signal devices, such as the 40673 and 3N211, are poor choices when high dynamic range is a design criterion. They are entirely acceptable for use in many low- and medium-cost hobby and entertainment receivers where high signal levels are not a problem. QST

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PERFORMANCE CAPABILITY OF ACTIVE MIXERS

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INTRODUCTION

Depending upon the application, there is a large variety of circuits used in passive and active mixers. It appears that mixers have a figure of merit expressed in the form of intermodulation distortion performance (intercept points of the order 1, 2, 3...N), suppression of harmonics and isolation, cut-off frequency, and LO drive.

The simple mixer consisting of one diode is only found in small pocket radios, and any high performance receiver or synthesizer application requiring mixers will make use of the harmonic canceling effect of double balanced mixers in a lattice configuration. Passive mixers have used either vacuum diodes, germanium diodes, silicon diodes or hot carrier diodes. The two basic requirements for these mixers are perfect match of the transformers and perfect match of the diodes. As the diodes are used in what is called "large signal application," the same non-linear performance of the transfer characteristic that is responsible for mixing generates harmonics of the input frequency and of the LO frequency, and these may appear at the output of the double balanced mixer if it is not carefully balanced. Perfect matching will prevent even order harmonics from appearing at the output, and so-called linear operation of the mixer, where the LO does not drive the non-linear device, will prevent excessive harmonic generation as such. Theoretically, mixers can be driven with square waves, another method of reducing harmonic combinations at the output.

While all passive mixers have losses, active mixers appear attractive because of their potential for showing gain.

Using active devices as mixers, we must consider three different applications:

a. Additive mixers

b. Multiplicative mixers

c. Switching operation, where the active device is used as a switch and operated without DC voltage.

From a device point of view, we have three different possibilities:

a. Bipolar transistors in mixers, which will result in additive mixing

b. Square law characteristic devices (junction field effect transistors, MOS field effect transistors, and enhancement field effect transistors (VMOS))

c. Dual gate MOSFETS, or IC type of mixers.

This paper will show some of the advantages, disadvantages, and high signal effects found in active mixers, their possible cures and trends. It should already be mentioned now that, for reasons explained very carefully in this paper, either the passive mixer with special diode ring configurations or the field effect transistors in a quad configuration used as a switch with no amplification is the ultimate choice for high performance. It has been shown only recently that intercept points of +40 dBm are possible using active devices in passive configurations with about 6 dB loss, 6 dB noise figure. Active mixers may be very useful in an environment with permanently constant amplitude like synthesizers; however, in the more hostile environment of receiver applications with today's technology, passive mixers are still less expensive, more reliable, and a higher performance circuit.

BASICS OF MIXERS

Mixing occurs in any non-linear device where the V/I curve deviates from a straight line if and when two or more signals will be applied to such a device.

The ideal and so-called linear mixer is a square law device like a field effect transistor with the transfer characteristic

$$i_D = I_{DSS}\left(1 - \frac{V_{GS}}{V_p}\right)^2 \qquad (1)$$

The transconductance is defined as the first derivative of d_i/d_v, and therefore, GM equals

$$GM = \frac{2I_{DSS}}{V_p^{\,2}}\,(V_p - V_{GS}) \qquad (2)$$

This is called linear mixing. It can be seen that the transconductance GM is a linear function of the gate source voltage V_{GS}.

Neglecting any non-linear effects like we would find in MOS field effect transistors or any reverse biasing effects as found in junction field effect transistors or inability to follow high frequency input voltage as found in VMOS transistors, the square law characteristic will only generate the second harmonic of the input and LO signal, and a perfect match in a balanced configuration would cancel this. The absence of a third order term would theoretically prevent any odd order intermodulation distortion product from happening. Such a square law characteristic is found in field effect transistors as mentioned, and for small signals, silicon or hot carrier diodes exhibit the same square law configuration.

A number of configurations are known using diodes in bridges to minimize harmonics at the output, and Figure 1a through 1f shows the series and shunt combination in which either two or four diodes can be used.

As shown in the literature (1), even with ideal diodes of zero forward resistance and infinite reverse resistance, the conversion loss of either the series or shunt modulator is π (9.9 dB). Practical modulators will have higher losses than this, as the diodes are not ideal.

Figure 2 shows the ring or lattice double balanced modulator as frequently used, and Figure 3 shows the latest two most important derivatives of the double balanced mixer, the two ring configuration and the termination insensitive mixer.

It has been explained very carefully in the literature (2) that all passive mixers are highly sensitive to changes in termination, and the reason for this is the non-zeroing effect of reactive currents at the output which generates reflections inside the bridge and, therefore, causes distortion.

Double balanced mixers are traditionally offered in 50Ω input and output impedance. In analyzing the functions of a double balanced mixer, it can be easily understood by replacing the diodes with switches that are being opened and closed as a function of LO drive. For reasons of available wire sizes and knowledge about building transmission lines with transformer wires, double balanced mixers are being built in a 50Ω input and output configuration. If the transformer was ideal and would not be a transmission line transformer, any output impedance would directly become visible at the input of the mixer. As most of RF applications now use 50Ω, this is very convenient. By using different wire sizes, the transformer can be changed, and as the cable television industry requires 75Ω mixers, this can be accomplished very easily. Additional external transformers can shift the impedance level to whatever is required. Figure 4 shows a mixer with additional balancing at the input and the output, as the assumption that the 4:1 or 1:4 transformer also provides ideal matching from unbalanced to balanced is not necessarily true. These discussions apply also to active mixers, as I have stated that the input and output port for the sake of suppression of harmonics must be balanced.

The best passive mixers show an intercept point of +30 to +35 dBm, use up to 64 monolithic diodes, and require up to +23 dBm LO. A push-pull configuration of two balanced mixers can show isolation of up to 60 dB over an extremely wide frequency range; the insertion loss is in the vicinity of 5.5 dB and then can be operated from 10 kHz to several GHz depending upon the transformers.

In the case of an active device, taking into consideration the linearities of the diode or active mixer, we can use the method of Fourier expansion to obtain the harmonic component of the LO pulse train of $0.2 = 2\pi/\omega$.

Figure 5 shows the train of sine wave tip current pulses if a sine wave, the LO, drives the slope of G that represents the transconductance. The resulting output can be used to determine the time average conductance of the device as a function of the conducting angle. In order to do this, we use the Fourier cosine expansion

$$f(t) = a_0 + a_1 \cos \omega t + a_2 \cos 2\omega t + \ldots$$

$$= a_0 + \sum_{n=1}^{\infty} a_n \cos n\omega t \qquad (3)$$

where

$$a_0 = \frac{1}{T} \int_{-T/2}^{T/2} f(t)\, dt \qquad (4)$$

and

$$a_n = \frac{2}{T} \int_{-T/2}^{T/2} f(t) \cos n\omega t\, dt \qquad (5)$$

By defining $\theta = \omega t$ and integrating over $d\theta$, we obtain

$$a_0 = \frac{1}{\pi} \int_0^{\pi} f\left(\frac{\theta}{\omega}\right) d\theta \qquad (6)$$

and

$$a_n = \frac{2}{\pi} \int_0^{\pi} f\left(\frac{\theta}{\omega}\right) \cos n\theta\, d\theta \qquad (7)$$

From Figure 5 it can be shown that the fundamental component

$$I_1 = \frac{2}{\pi} \int_0^{\phi} G(V_1 \cos \theta - V_x) \cos \theta d\theta$$

$$= \frac{2G}{\pi} \left(\frac{V_1 \phi}{2} + \frac{V_1 \sin 2\phi}{4} - V_x \sin \phi \right)$$

$$= \frac{I_p}{\pi} \frac{\phi - \cos \phi \sin \phi}{1 - \cos \phi} \qquad (8)$$

In a similar way, we obtain

$$I_0 = \frac{I_p}{\pi} \frac{\sin \phi - \phi \cos \phi}{1 - \cos \phi} \qquad (9)$$

and

$$I_n = \frac{2I_p}{\pi} \frac{\cos \phi \sin n\phi - n \sin \phi \cos n\phi}{n(n^2 - 1)(1 - \cos \phi)}, \quad n \geq 2 \qquad (10)$$

As explained in my previous paper (3), Figure 5 can be drawn plotting the normalized output, normalized voltage gain, and normalized mixing transconductance S as a function of normalized oscillator voltage. From Figure 6, we would see a practical value for x = 0.75, and we get a mixing transconductance $G_m = 0.56 \cdot G_M = 2.5$ mS for a 2N3822 field effect transistor. For a higher order transfer characteristic, the approach would be the same, and the equation for I as a function of V would change.

As mentioned previously, we have three types of mixing.

a. Additive mixing. Additive mixing is based upon the fact that the two components $v_1(t) + v_2(t)$ can be rewritten in the form

$$V = V_1 \cos \omega_1 t + V_2 \cos \omega_2 t \qquad (11)$$

The expansion of this leads to the product

$$C(t) \{\cos (A - B)t + \cos (A + B)t\}$$

Additive mixing would occur where the two signals are being fed in series. All field effect and bipolar transistors where the LO is applied either together with the RF signal to the same electrode (gate, base, source, or emitter) or whether the two signals are applied to different electrodes, use the additive principle.

b. Multiplicative mixing. Only in the case of the dual gate MOSFET and in the case of a differential amplifier with a constant current source can we use the term "multiplicative mixing." However, the net result remains the same. The advantage in using multiplicative mixers is that the schematics provide isolation between the two ports, which means that there is very little or no interaction between the RF and the LO port.

Figure 7 shows a recommended circuit for the Motorola MC1596 integrated circuit which is the basis for the Plessey mixer type SL6440 shown in its test circuit, Figure 8. Plessey reports an intercept point in the vicinity of +30 dBm, about 0 dB gain, and roughly 11 dB noise figure.

c. Mixing by switching. In the case of the double balanced mixer with diodes, the diodes act as a switch. It must be assumed that these switches are fast enough to follow the local oscillator, and therefore, hot carrier diodes are found for high frequency operations. Because of the switching, the input and output impedances are reflected at the output and input, and the mixer becomes transparent. The insertion loss is primarily determined by the fact that the sum and difference of the two signals is at the output, and only one of them is the wanted signal. If the input voltage is divided into two output voltages, we must have 3 dB losses. The additional losses occur from the effect that the diodes have series resistors which are responsible for the losses. The amount of resistive loss here is in the vicinity of 2% to 3% due to the 1Ω the diodes exhibit under switched-on conditions. Ideally, this type of mixing does not depend upon any transfer characteristic, and we will see later that if this type of operation is duplicated with active devices, we will obtain the best possible performance.

SIGNAL HANDLING

The characteristic of the nonlinear device again can be expanded in the form

$$g_m = a_{01} + \frac{a_{02}}{2!} v + \frac{a_{03}}{3!} v^2 + \frac{a_{04}}{4!} v^3 + \dots$$
$$+\left(a_{11} + \frac{a_{12}}{2!} v + \frac{a_{13}}{3!} v^2 + \frac{a_{14}}{4!} v^3 + \dots\right) \cos \omega_0 t$$
$$+\left(a_{21} + \frac{a_{22}}{2!} v + \frac{a_{23}}{3!} v^2 + \frac{a_{24}}{4!} v^3 + \dots\right) \cos 2\omega_0 t$$
$$+ \dots \qquad (12)$$

The following significant interfering effects can be distinguished:

a. Hum modulation, expressed by

$$m_u \simeq \frac{a_{12}}{a_{11}} V_u \qquad (13)$$

where m_u = undesired modulation of carrier, and V_u = amplitude of a-f voltage causing modulation.

b. Variation of the modulation depth, expressed by

$$M \simeq \frac{\Delta m}{m} = \frac{1}{4}\left(\frac{a_{13}}{a_{11}}\right) V_1^2 \qquad (14)$$

where V_1 = average amplitude of desired signal.

c. Modulation distortion, expressed by

$$D_2 \simeq \frac{3}{16}\left(\frac{a_{13}}{a_{11}}\right) V_1^2 \qquad (15)$$

where V_1 = average amplitude of desired signal.

d. Cross-modulation, expressed by

$$K = \frac{m_k}{m} \simeq \frac{1}{2}\left(\frac{a_{13}}{a_{11}}\right) V_u^2 \qquad (16)$$

where V_u = average amplitude of undesired signal.

e. Spurious responses at $n_1 = 1$, $n_0 = x$, expressed by

$$\frac{V_1}{V_u(x,1)} \simeq \frac{a_{x1}}{a_{11}} \qquad (17)$$

where V_1 = average amplitude of desired signal, and $V_u(x,1)$ = amplitude of spurious signal giving the same output as the desired signal.

f. Spurious responses at $n_1 = 2$, $n_0 = x$, expressed by

$$\frac{V_1}{V_u(x,2)} \simeq \frac{a_{x1}}{4a_{11}} V_u(x,2) \qquad (18)$$

where V_1 = average amplitude of desired signal, and $V_u(x,2)$ = amplitude of spurious signal giving the same output as the desired signal.

The coefficients of Equation 12 depend on the $i_2 = f(v_1, v_0)$ characteristics of the mixer. If, for example, the pseudo-static current I_2 of an additive mixer is known as a power series, $I_2 = I_2(0) + p_V + q_V{}^2 + r_V{}^3 + s_V{}^4 + t_V{}^5 + \dots$ then for $V \to v + V_0 \cos \omega_0 t$, $I_2 \to i_2$, and since $i_2 - I_2(0) = g_m(t)v$,

$$a_{01} \simeq p + 3/2 rV_0{}^2 + \dots \qquad (19)$$

$$\frac{a_{02}}{2} \simeq q + 3sV_0^2 + \ldots \qquad (20)$$

$$\frac{a_{03}}{6} \approx r + 5tV_0^2 + \ldots \qquad (21)$$

$$a_{11} \approx 2qV_0 + 3sV_0^3 + \ldots \qquad (22)$$

$$\frac{a_{12}}{2} \approx 3rV_0 + 15/2tV_0^3 + \ldots \qquad (23)$$

$$\frac{a_{13}}{6} \approx 4sV_0 + 15uV_0^3 + \ldots \qquad (24)$$

$$a_{21} \approx 3/2rV_0^2 + 5/2tV_0^4 + \ldots \qquad (25)$$

$$\frac{a_{22}}{2} \approx 3sV_0^2 + 15/2uV_0^4 + \ldots \qquad (26)$$

The coefficients depend on the bias point. Using theoretical characteristics of the various mixers often leads to inaccurate results because the influence of parasitic effects may be considerable.

PRACTICAL CIRCUITS

For reasons of balancing harmonics and other unwanted products, we will not consider simple single stage mixers.

Figure 9 shows the schematic of an active mixer using two VHF field effect transistors in push-pull. Because of the number of spurious still generated, substantial filtering at the output is recommended, as can be seen from the high pass/low pass filter section. The two RF chokes at the base of each transistor prevent unwanted oscillation at fairly high frequencies.

Figure 10 shows a balanced mixer using the 3N200 field effect transistor in push-pull. This is a multiplicative mixer where the RF is fed in push-pull, and the LO is fed in parallel.

Ed Oxner of Siliconix has designed several mixers based on the U257 transistor. The schematics are shown in Figure 11. Until recently, Siliconix made a version with four field effect transistors in one case, which now is discontinued. It appears that there was not enough demand for the quad mixer. Figure 12 shows the schematic of such a configuration, using the U350 which is no longer available.

VMOS transistors have become popular and a push-pull version with the VMP4 power FET as described by Doug DeMaw was shown in QST, January 1981, (Figure 13).

This collection of circuits pretty much represents the state-of-the-art using active mixers, and depending upon bias impedance ratio at device selection, fairly high intercept points up to +30 dBm were obtained.

Most loads in which the mixers operate do not present precisely 50Ω. Most designers of active mixers have carefully avoided indicating the effects that happen in active mixers as the termination changes from purely resistive 50Ω into something. In general, any change in resistive load without introducing reactive components does not affect the mixer substantially. A VSWR of 1:2 from 50Ω, or change of load from 25Ω to 100Ω resistive does not have too many adverse effects. If, however, the mixer is terminated by an LC filter or crystal filter, that changes the impedance, and becoming reactive, the intercept point changes drastically. In some cases with active mixers, I have observed even instabilities to the point where the mixer turned into a low frequency oscillator.

There are three basic circuits known to prevent change of impedance. Figure 14 shows a recommended arrangement whereby the mixer, in this case a passive double balanced mixer, is terminated by the input impedance of a grounded gate field effect transistor. It must be remembered that grounded gate field effect transistors properly biased exhibit purely resistive input over an extremely wide frequency range. This holds true in most cases in basically DC from several hundred MHz. The CP643 or CP640 made by Teledyne Crystalonics is a good choice.

Another alternative is a feedback amplifier that uses noiseless feedback as described in the literature (5,6) based upon Patent No. 3891934 of 1975. The third alternative is the use of a diplexer whereby the image at the output of the mixer is terminated into a 50Ω resistor.

Probably the best solution is a combination of two and three as shown in Figure 15. Again, for reasons of convenience, the circuit is shown with a passive balanced mixer together with this particular termination circuit.

Let us now take a look at some systems calculations which will yield a surprising result.

a. Active mixer with perfect termination. Consider an active mixer like the Plessey SL6440 on any of the previously shown schematics. The noise figure under large signal operation is around 11 dB for the Plessey device and 8 dB for a U257 mixer.

Relative to the typical loss of 6 dB in a passive mixer, the 0 dB gain of an active mixer already represents gain, to be specific 6 dB gain over the passive device. Let us assume further that the following amplifier uses the noiseless feedback and its noise figure is 2 dB. As the mixer has unity gain, the noise figure at the input is equal to the noise figure of the second stage plus the noise figure of the mixer, and in the case of the Plessey mixer, the resulting noise is 13 dB.

If we use a U257 stage, we get a 10 dB noise figure if we allow the same amount of gain. The intercept point is determined by the mixer and the second amplifier, and because of the special RF feedback applied in the second amplifier, we will, for the moment, assume that the second amplifier does not contribute any intermodulation distortion products. The very moment we operate the mixer with gain, we have to take distortion of the second stage into consideration.

b. Passive double balanced mixer with termination stage. Let us use the same example with a high performance double balanced mixer. The double balanced mixer has an insertion loss of 6 dB, and the noise figure also is 6 dB.

The noise figure of the termination stage again is 2 dB, which results in a total systems noise figure of 8 dB, or 2 dB better than the previous example with the U257. Because of the 6 dB losses of the double balanced mixer, the intermodulation distortion of the double balanced mixer can be neglected, and the analysis can be concentrated on the mixer itself. This, I am sure, is a surprising result for most design engineers.

It is important to understand that the terminating stage, when using the noiseless feedback system, has to operate into a stable load also. Any changes of the output load of such an RF feedback amplifier will be reflected into the input. A recommended way of reducing this is to operate this stage at higher than necessary gain; 3 or 4 dB is sufficient. A resistive pad with 3 dB attenuation then will prevent dramatic changes at the output.

In the case of the field effect transistor and grounded gate as termination, this circuit only works reliably if the feedback capacitor C_{DS} is kept extremely small and is basically determined by the transistor itself.

If the output stage has to operate into a crystal filter, we will find that most crystal filters outside the passband characteristic become high impedance and either inductive or capacitive. The effect of this can be reduced by using a high pass filter at the output that incorporates the crystal filter. If the crystal filter impedance increases, the high pass filter is mistuned and the voltage at the drain or collector remains low. As a result, the third order intermodulation distortion products remain low. In a conventional circuit, it is sometimes found the sudden increase of impedance at the output of the transistor makes the intermodulation distortion deteriorate.

Passive Mixer with Active Devices

Recent development work in the field of mixers indicates that the best way of achieving high intercept points in mixers is the use of

a. Bipolar transistors as switches and with feedback

b. Field effect transistors as switches.

In ordinary applications, active devices operate based upon the nonlinear transfer characteristic as explained earlier. Diode mixers are substantially better because here the device is only switched on and off, and if the on/off resistance of the device has a high enough ratio, the device is fast enough to follow the LO drive waveform, and enough LO drive power is available, theoretically, no harmonic distortion would occur. The losses would be 3 dB, and so would be the noise figure, and we would not observe any intermodulation distortion products at all. Using active devices, we depend upon the non-linearities of the input and output ports and like in the field effect transistor possible distortion of the gate source diode and the potential non-linearities of the channel resistance.

The state-of-the-art in mixers using field effect transistors without operating voltage and, therefore, only as switches is about +42 dBm intercept point, 5.5 dB noise figure and insertion loss, and LO requirements of about +23 dBm. The LO drive requirement really results from the effect that a certain voltage has to be available at the gate electrode. In a quad configuration, this voltage can be as high as 50V PEP into the input capacitance of the transistor. The step-up transformer helps to reduce the required power.

Figure 16 shows the schematic of such a recommended mixer which, for test purposes, has a tuned input. This circuit is based upon a patent issued to Mr. William Squires in 1968, No. 3383601. It can be reported that for 1V input signal or +13 dBm, the third order intermodulation distortion products are at -83 dBm or 100 dB down. This would increase to an intercept point of +70 dBm, but this can only be achieved in narrowband circuit. In a wideband configuration, only 40 to 42 dBm is obtainable. The isolation between oscillator and signal port is about 60 dB and provides about 40 dB isolation to the IF.

This area of using passive mixers with active devices is fairly new. The only company that seems to have a commercial product is Lorch in New Jersey, and the latest prices I have seen for their mixer were $600 or $700.

I had built an active mixer based upon feedback and switching which was published in Ham Radio Magazine (7). This mixer with similar performance was used in the Rohde & Schwarz 400W transceiver in the Tornado warplane.

TESTING

In order to make proper tests on the mixers to signal generators, a hybrid coupler with at least 40 dB isolation between the two input ports and an attenuator is required. The test set-up provided by DeMaw in QST, January 1981 (4) and shown in Figure 17 is ideal for this. He has taken two signal generators around 14 MHz, combined them, and while an attenuator drives the mixer (mixer under test = MUT), the LO is supplied by a VFO, and the output then is analyzed.

The 2N5109 amplifier shown may not be sufficient for extremely high intercept points as this stage may no longer be transparent, and for reasons of stability tests when using active mixers, it is recommended to have a reactive network at the output of the mixer for the sole purpose of checking whether the mixer can become unstable.

The two oscillators have to have extremely low harmonic contents and very low noise sideband performance. A convenient circuit to provide the required harmonic suppression and low noise is shown in Figure 18 based upon an earlier paper of mine (8). For those who are interested in having an additional 20 dB improvement in noise sideband and need a test oscillator of this performance, the circuit, Figure 19, is recommended. This oscillator shows an ultimate noise floor of 168 dB/Hz at 1 kHz off the carrier. As it can be seen, this oscillator is a derivative of the earlier one. The input impedance of the grounded base stage is about 2Ω and, therefore, does not really deteriorate the Q of the crystal.

SUMMARY

It has been explained that ordinary active mixers based upon the inherent non-linearities of their transfer characteristic by definition will show a lower intercept point than is possible with passive devices. Passive devices are already used with great success, and the termination insensitive mixers, although they are not yet offered below 1 MHz, are currently in the state-of-the-art in diode mixers. By either using feedback techniques together with switching type active stages and bipolar transistors or better using modern power junction field effect transistors, intercept points to 40 dBm third order and higher order are possible. In selective cases and narrow frequency operation, +70 dBm intercept points have been reported. It is not very likely that these figures are useful as the termination stages or following crystal filters or other devices will become the limiting factor.

I have just learned that Mini-Circuit Laboratories came out with a new mixer type VAY1 that claims 38 dB intercept point at the input, which results in 32 dBm intercept point at the output. However, the drive requirements are much higher than for the passive FET mixer.

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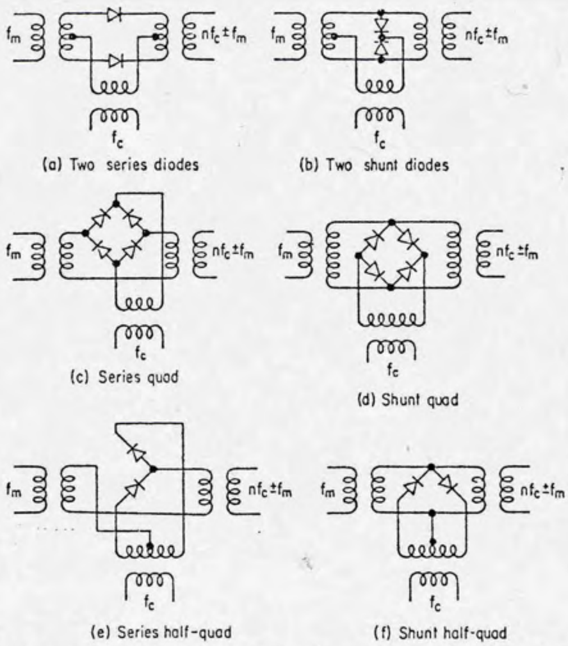

Figure 1 - Series and shunt type of mixers

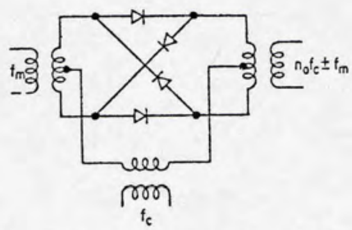

Figure 2 - Standard level double balanced mixer

Figure 3a and b - High level double balanced mixer

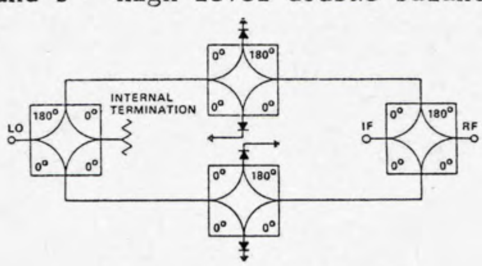

Figure 3c - Termination insensitive mixer

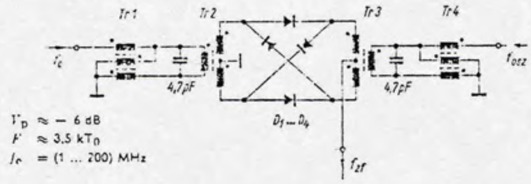

Figure 4 - Practical circuit for a double balanced mixer including input and output balancing transformers

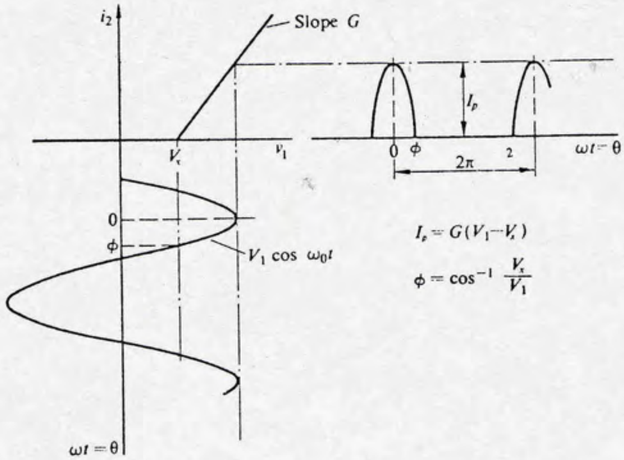

Figure 5 - Sine wave tips representing the time variable transconductance of a square wave transfer characteristic device

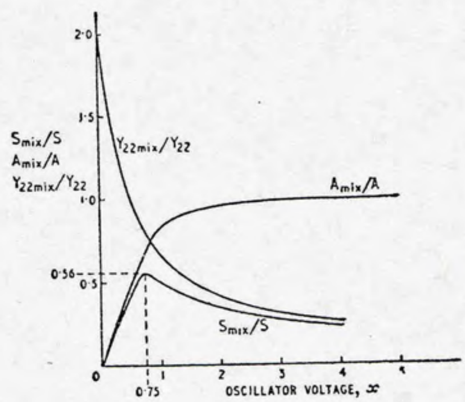

Figure 6 - Normalized voltage gain output impedance and mixing transconductance (s) for the FET

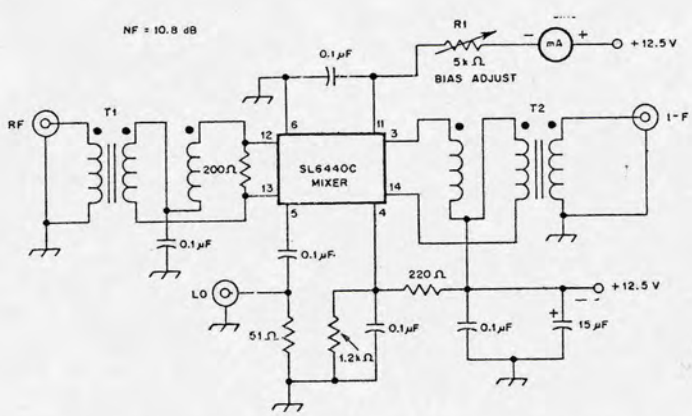

Figure 7 - Recommended test circuit for the SL6440 active double balanced mixer

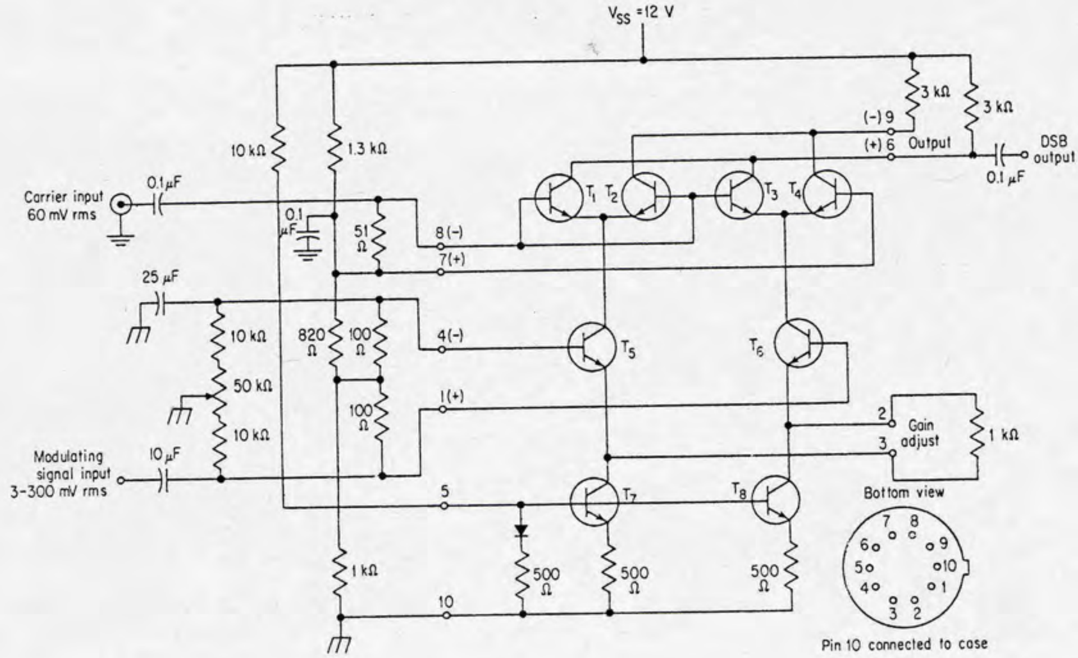

Figure 8 - Active mixer using the Motorola MC1596

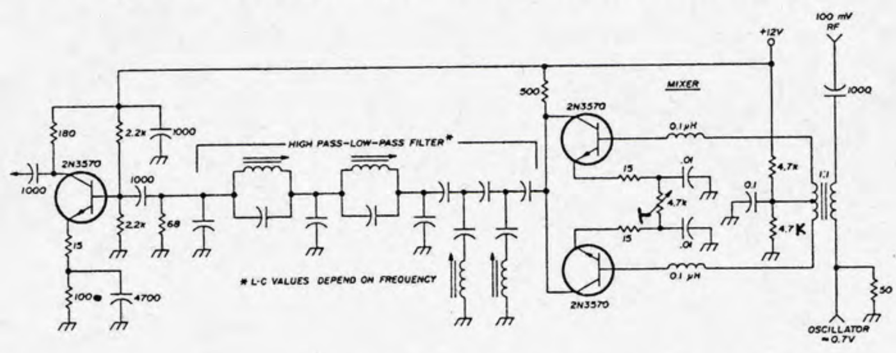

Figure 9 - Active mixer with transistors and bandpass filter

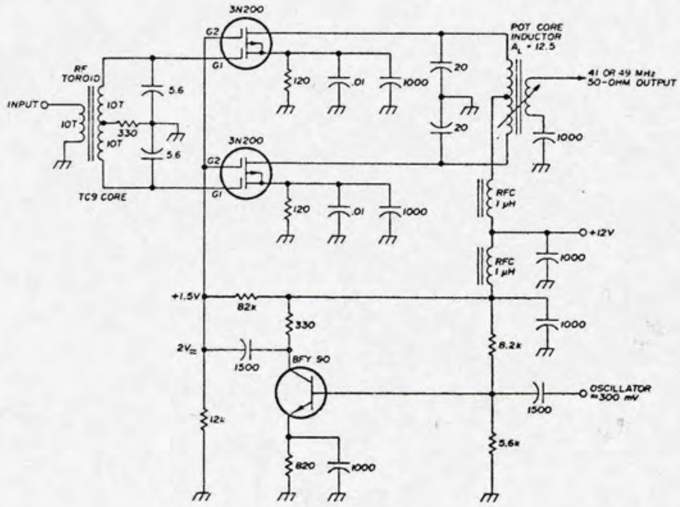

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Figure 10 - Push-pull mixer with 3N200 FET

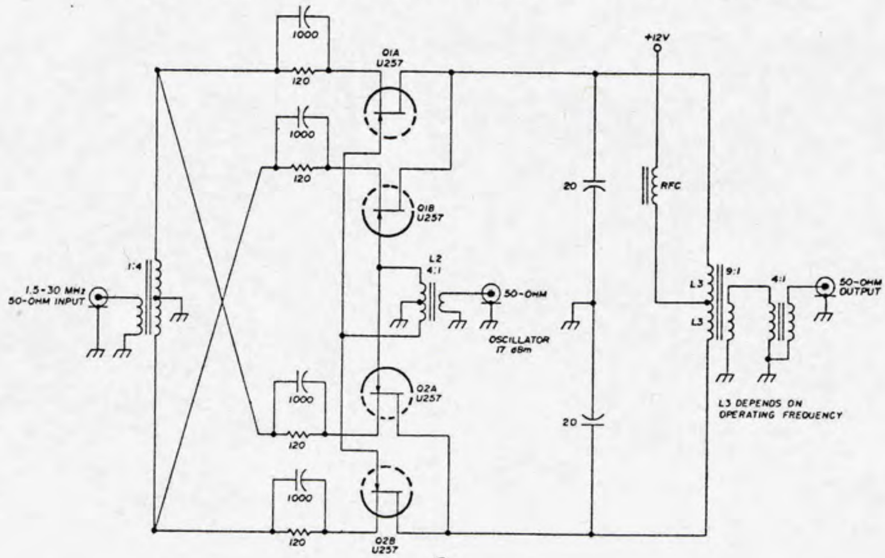

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Figure 11 - Double balanced mixer with Siliconix U257

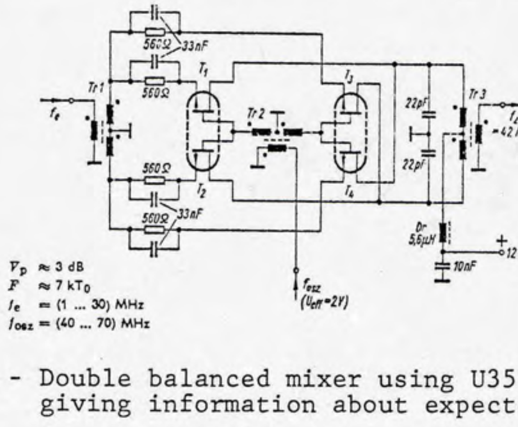

Figure 12 - Double balanced mixer using U350 transistor giving information about expected performance

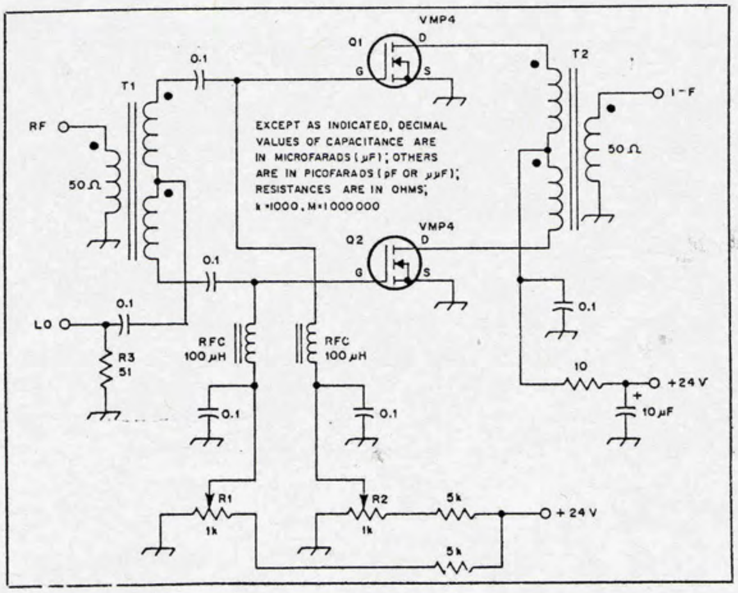

Figure 13 - Push-pull mixer with VMP4 FET's

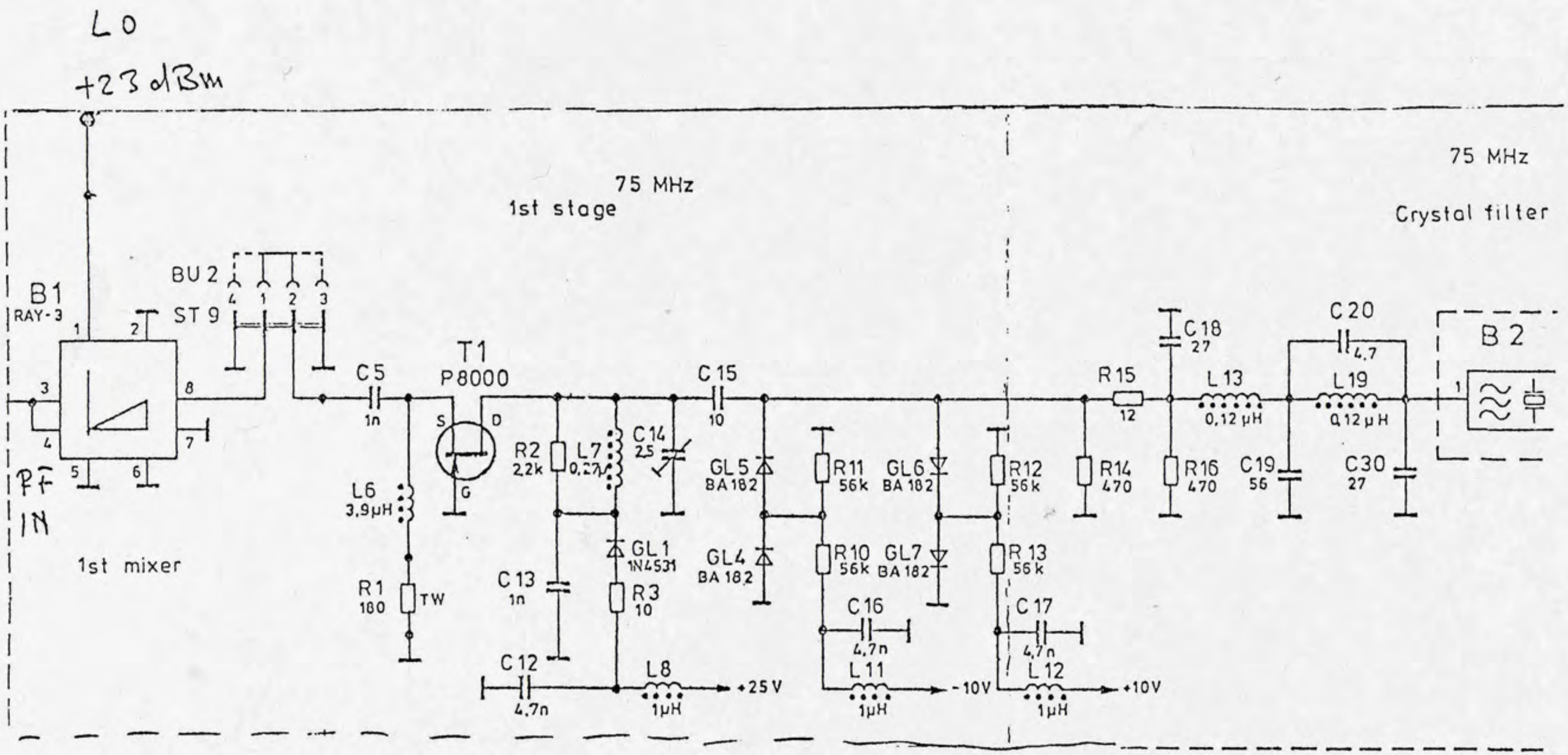

Figure 14 - Double balanced mixer with grounded gate FET as post-amplifier

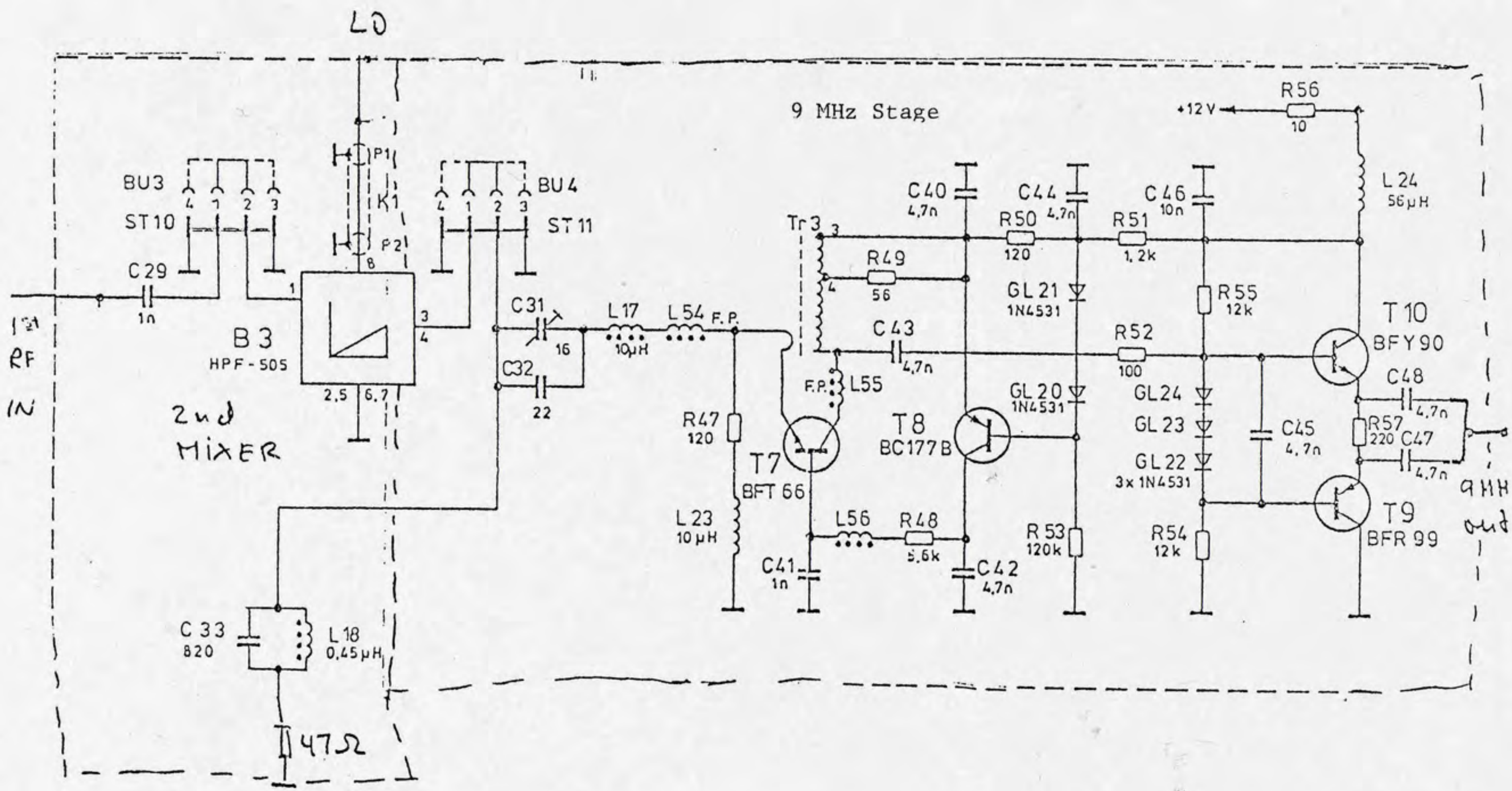

Figure 15 - Double balanced mixer with noiseless feedback amplifier as well as diplexer as termination

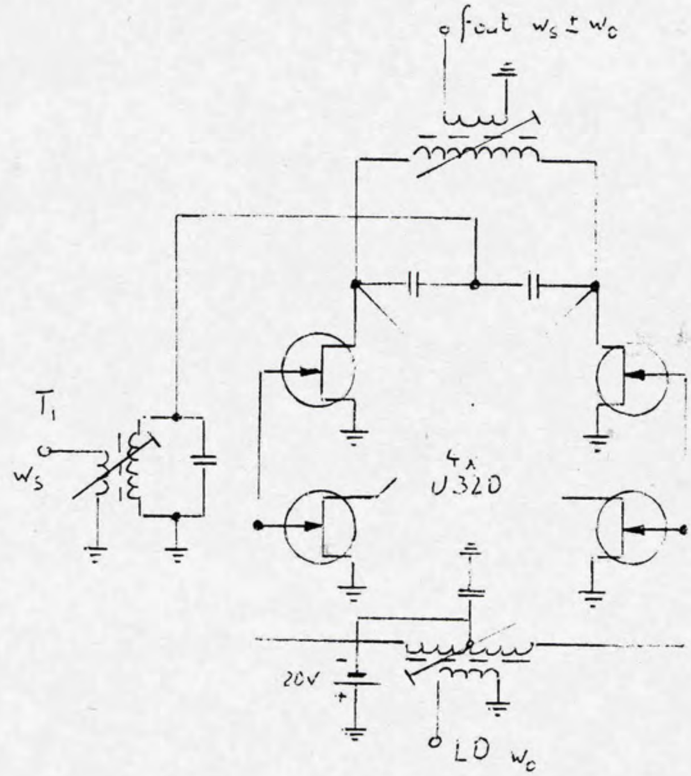

Figure 16 - Schematic of a passive double balanced mixer using FET's in a quad arrangement. This circuit represents the state-of-the-art that is possible today. While the narrowband version can have input intercept points of +70 dBm, a wideband version achieves about +42 dBm.

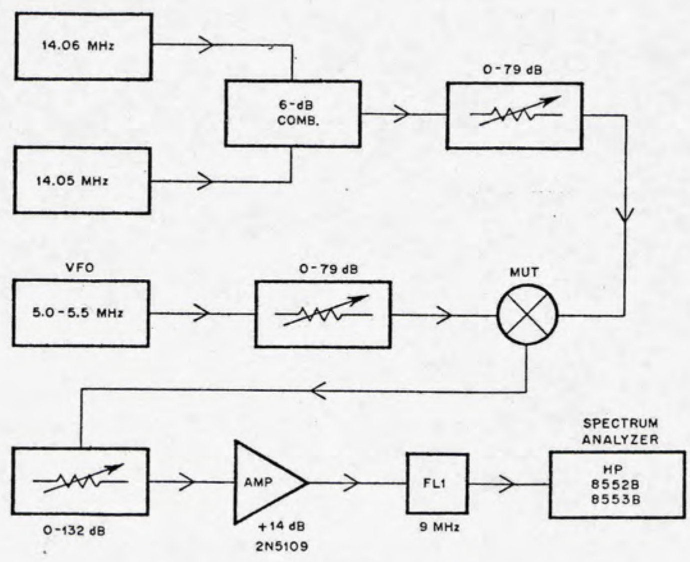

Figure 17 - Recommended test set-up for measuring mixer intermodulation distortion

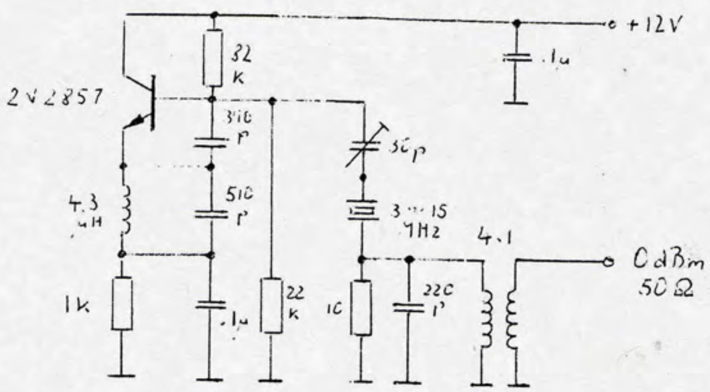

Figure 18 - Low noise crystal oscillator with
60 dB harmonic suppression

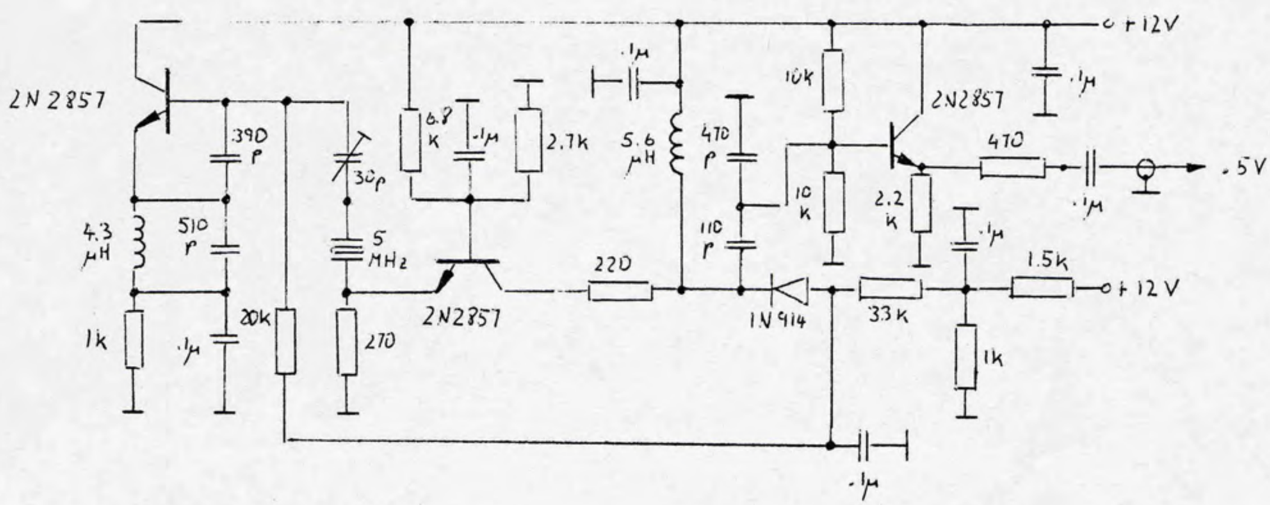

Figure 19 - Ultra low noise crystal oscillator with ultimate
noise floor or 168 dB/Hz