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

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

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

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

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

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

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. <u>Additive mixing</u>. 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. <u>Multiplicative mixing</u>. 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. <u>Mixing by switching</u>. 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 non-linear device again can be expanded in the form

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

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, $I_2 = I_2(0) + p_V + q_V{}^2 + r_V{}^3 + s_V{}^4 + t_V{}^5 + \ldots$ then for $V \to v + V_0 \cos \omega_0 t$, $I_2 \to i_2$, and since $i_2 - I_2(0) = g_m(t)v$,

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

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

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

$$a_{11} \approx 2qV_0 + 3sV_0^3 + \ldots \tag{22}$$

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

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

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

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

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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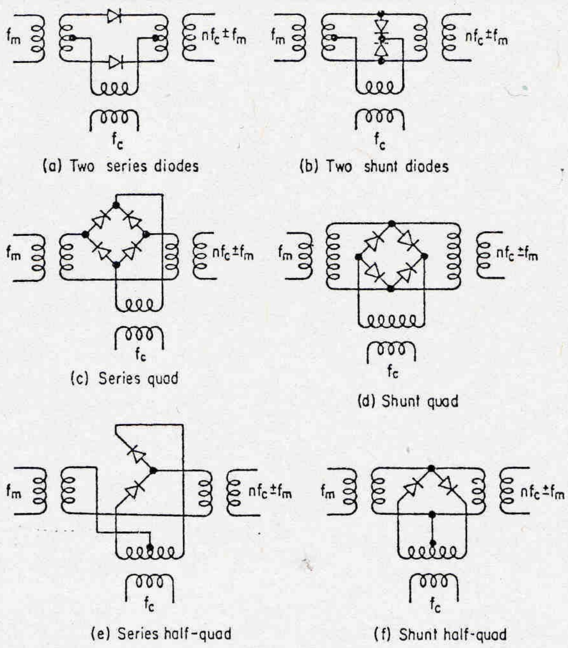

Figure 1 - Series and shunt type of mixers

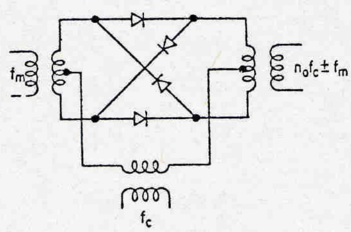

Figure 2 - Standard level double balanced mixer

Figure 3a and b - High level double balanced mixer

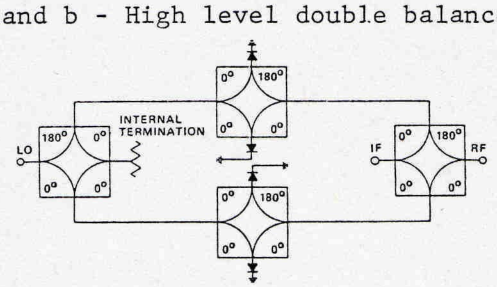

Figure 3c - Termination insensitive mixer

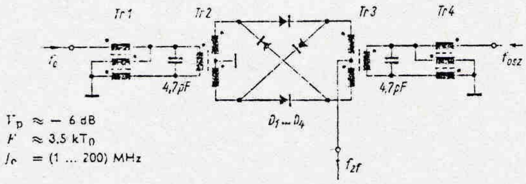

Figure 4 - Practical circuit for a double balanced mixer including input and output balancing transformers

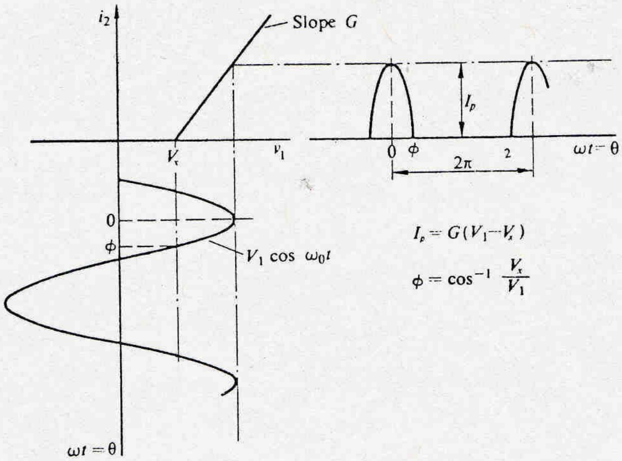

Figure 5 - Sine wave tips representing the time variable transconductance of a square wave transfer characteristic device

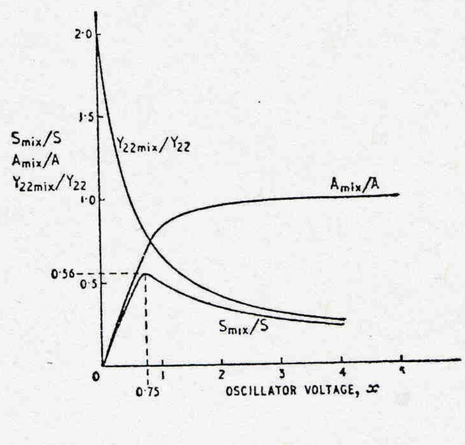

Figure 6 - Normalized voltage gain output impedance and mixing transconductance (s) for the FET

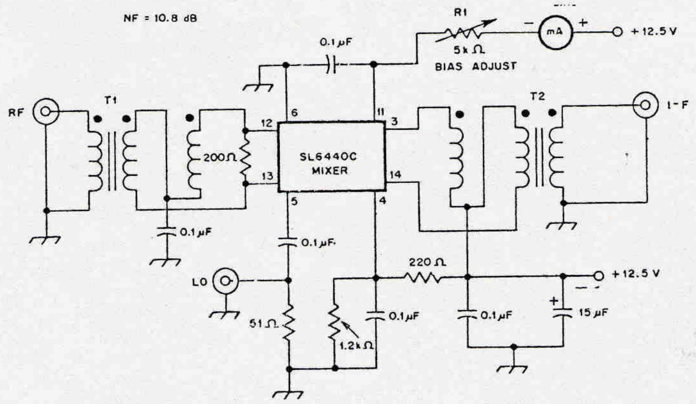

Figure 7 - Recommended test circuit for the SL6440 active double balanced mixer

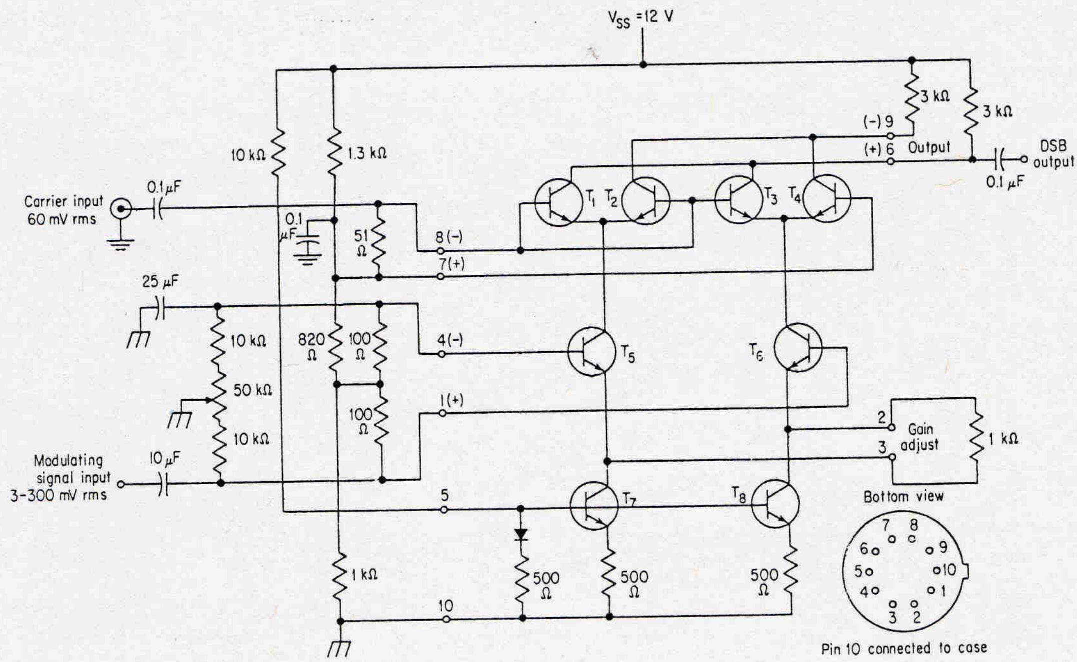

Figure 8 - Active mixer using the Motorola MC1596

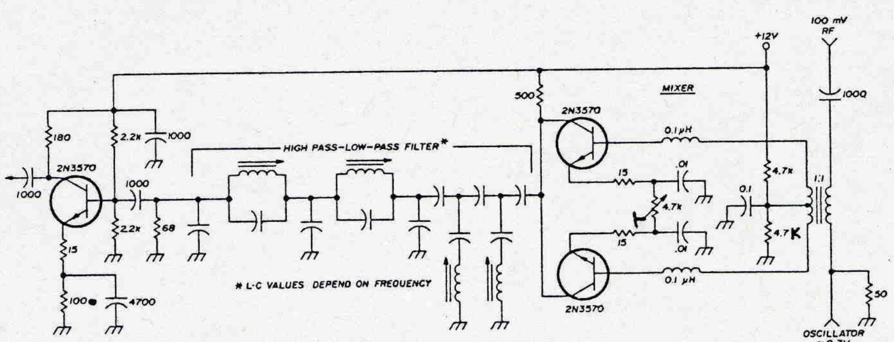

Figure 9 - Active mixer with transistors and bandpass filter

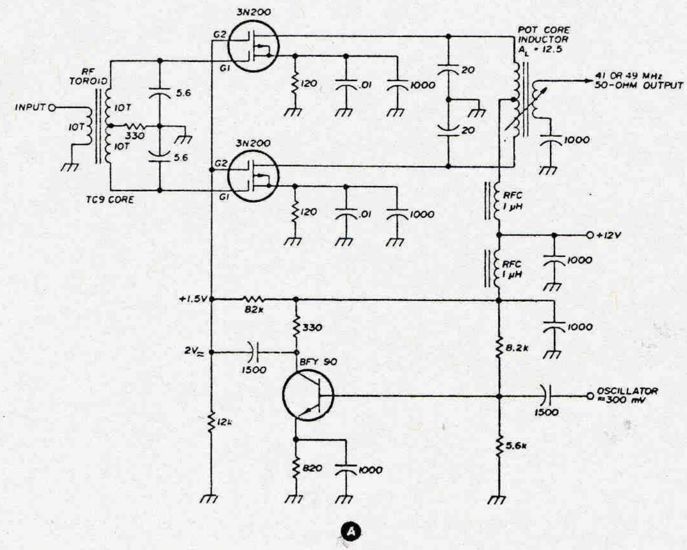

Figure 10 - Push-pull mixer with 3N200 FET

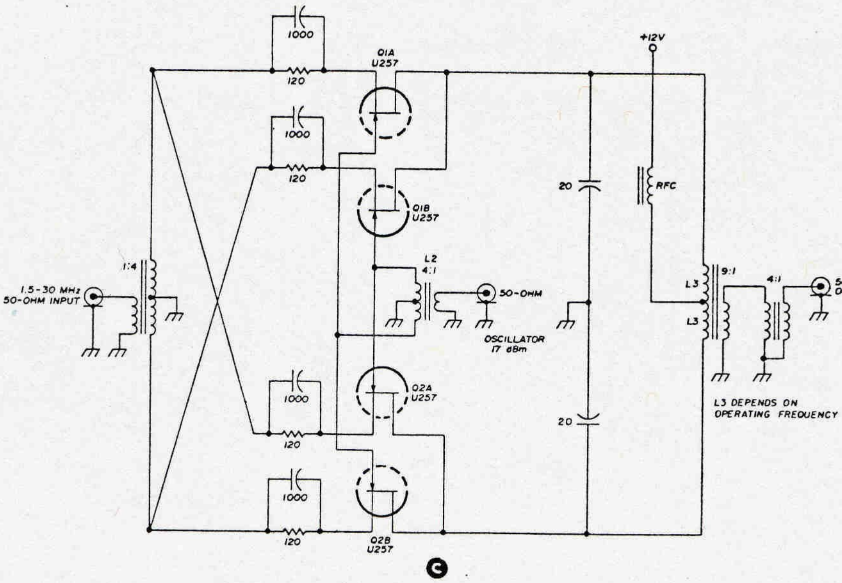

Figure 11 - Double balanced mixer with Siliconix U257

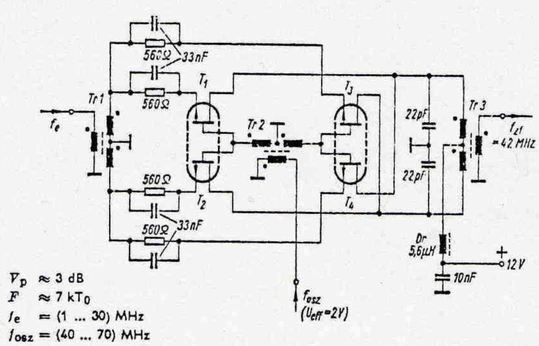

Figure 12 - Double balanced mixer using U350 transistor giving information about expected performance

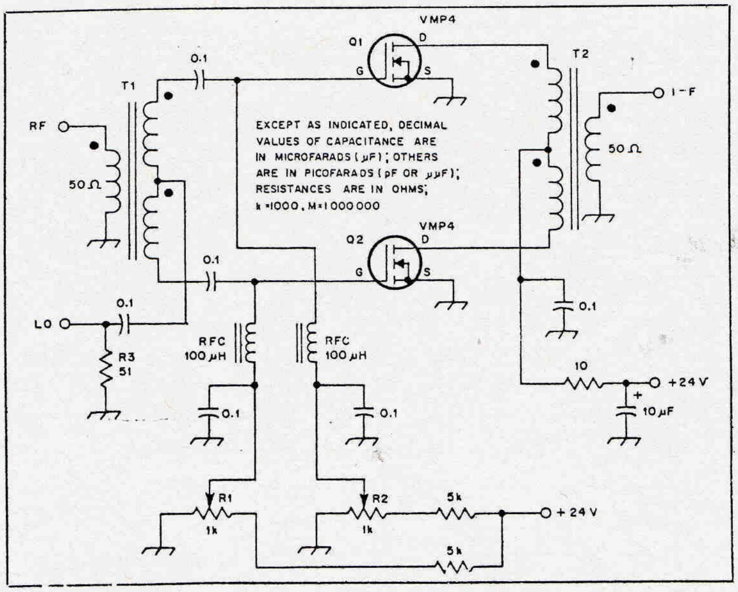

Figure 13 - Push-pull mixer with VMP4 FET's

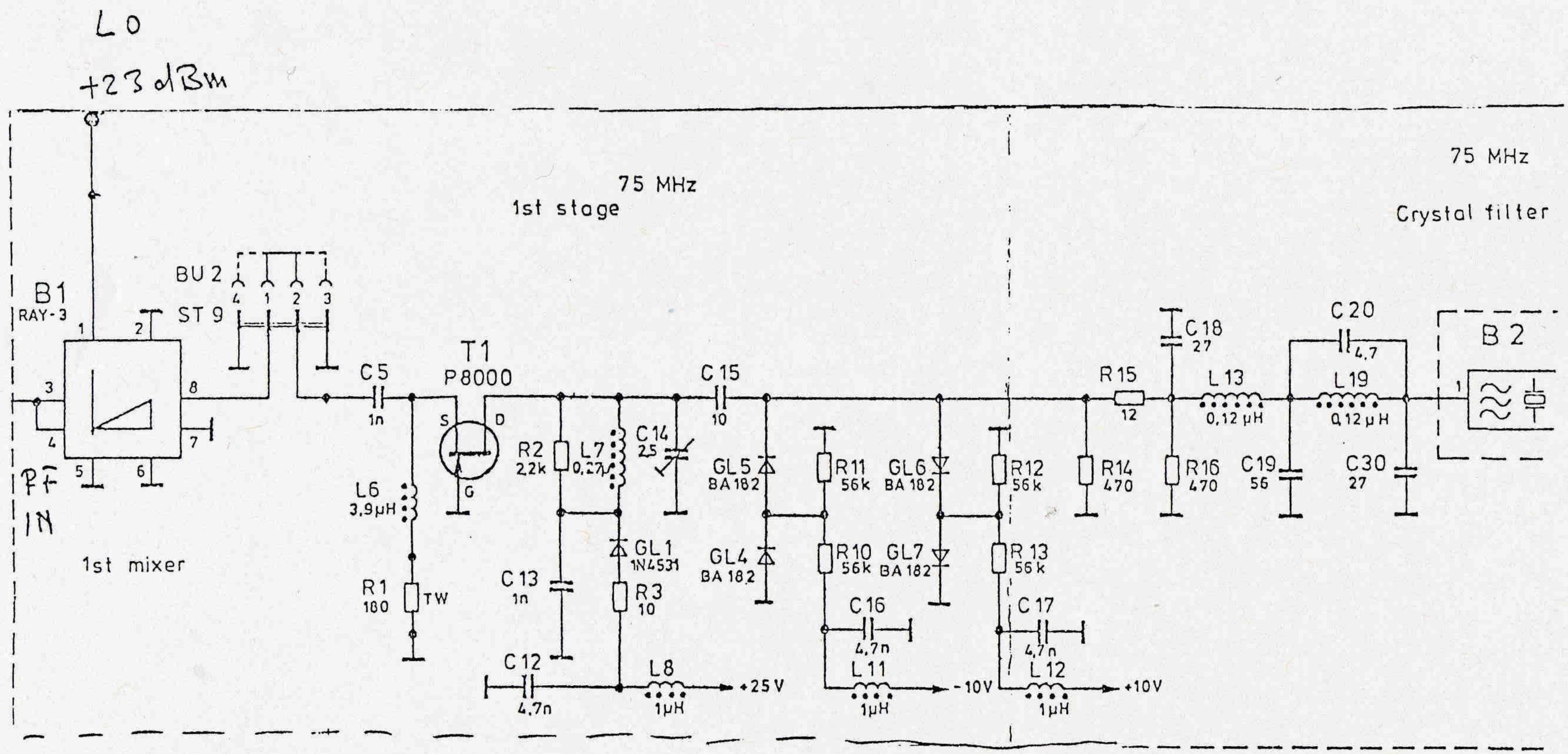

Figure 14 - Double balanced mixer with grounded gate FET as post-amplifier

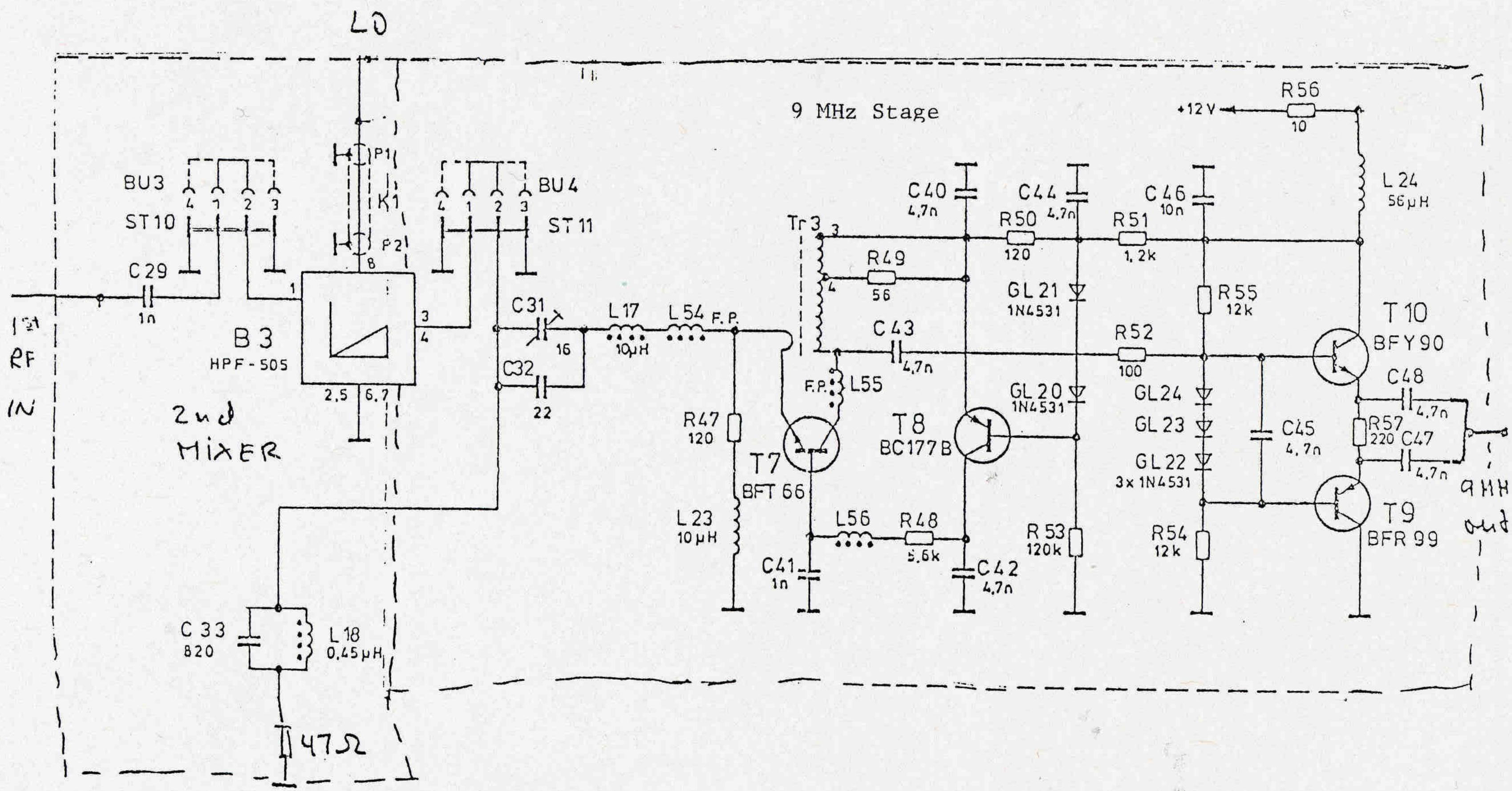

Figure 15 - Double balanced mixer with noiseless feedback amplifier as well as diplexer as termination

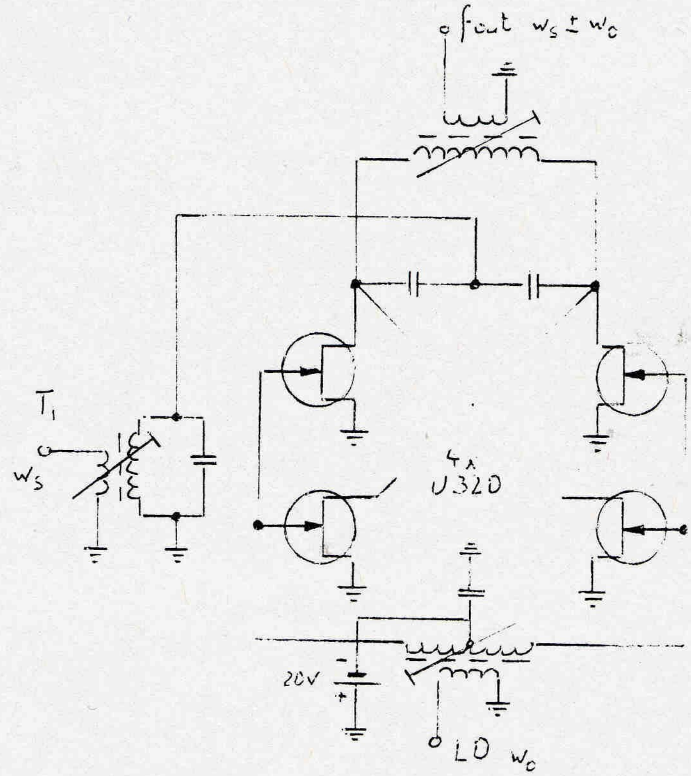

Figure 16 - Schematic of a passive double balanced mixer using FET's in a quad arrangement. This circuit represents the state-of-the-art that is possible today. While the narrowband version can have input intercept points of +70 dBm, a wideband version achieves about +42 dBm.

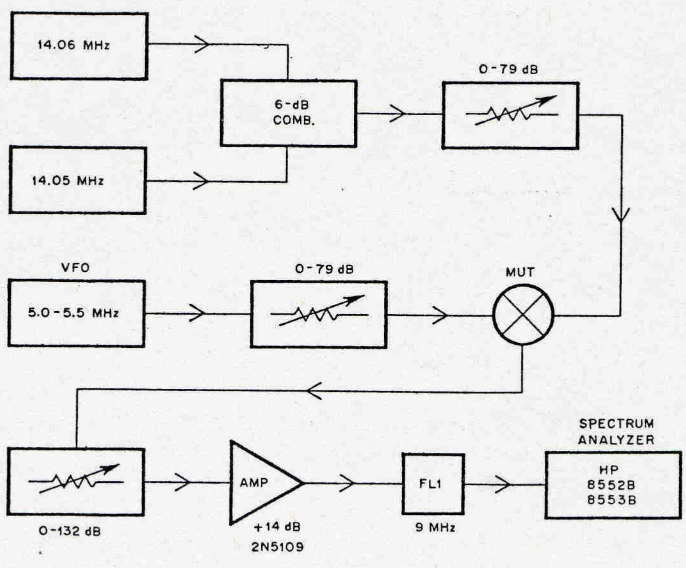

Figure 17 - Recommended test set-up for measuring mixer intermodulation distortion

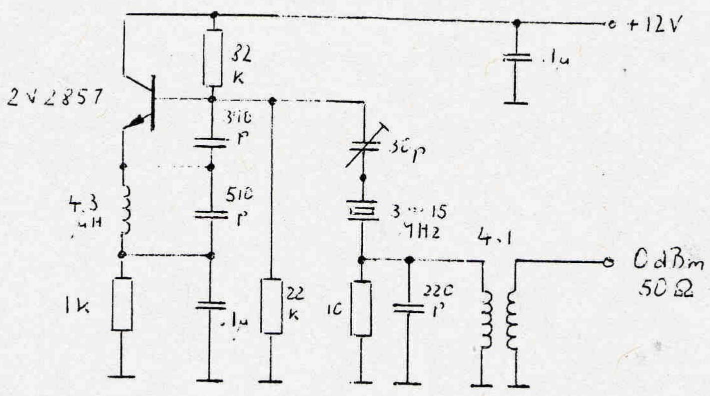

Figure 18 - Low noise crystal oscillator with
60 dB harmonic suppression

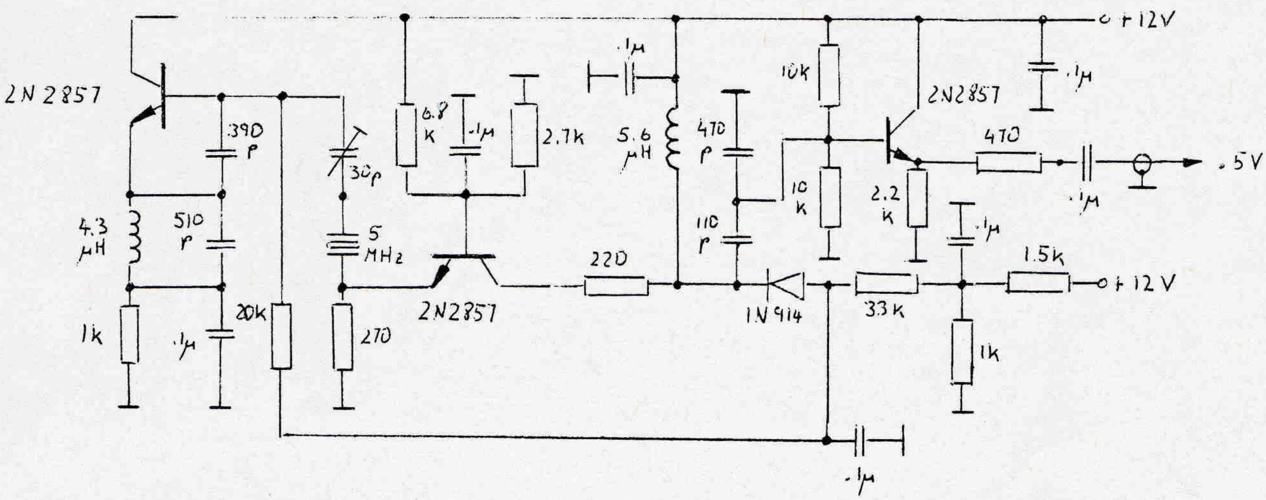

Figure 19 - Ultra low noise crystal oscillator with ultimate
noise floor or 168 dB/Hz