RF mixing

Having examined the basics of rf mixer operation, and demonstrated the basic single-ended diode mixer, consultant Joe Carr now looks at a few of the more important performance parameters, and a circuit or two. First, mixer distortion products...

Because mixers are non-linear, they will produce both harmonic distortion products and intermodulation products Our main interest at this point is the intermodulation products, which from hereon, I will shorten to IPs

Intermodulation products

The spurious IP signals generated when two signals, F_1 and F_2, are mixed non-linearly are shown graphically in **Fig. 1**, assuming input frequencies of 1MHz and 2MHz Given input signal frequencies of F_1 and F_2, the main IPs are:

Second-order:	$F_1 \pm F_2$
Third-order:	$2F_1 \pm F_2$ $2F_2 \pm F_1$
Fifth-order:	$3F_1 \pm 2F_2$ $3F_2 \pm 2F_1$

The second-order and third-order products are those normally specified in a receiver mixer design because they tend to be the strongest.

In general, even-order intermodulation distortion products, 2, 4, etc., tend to be less of a problem because they can often be ameliorated by using external filtering ahead of the receiver mixer – or tuned rf amplifier if one is used. Pre-filtering tends to reduce the amplitude of out-of-channel interfering signals, reducing the second-order products within the channel.

Third-order IM distortion products are more important because they tend to reflect on the receiver's dynamic range, and its ability to handle strong signals On the whole, the third-order products are not easily influenced by external filtering, so must be handled by proper mixer selection and/or design

When an amplifier or mixer is overdriven, the second-order content of the output signal increases as the square of the input signal level, while the third order responses increase as

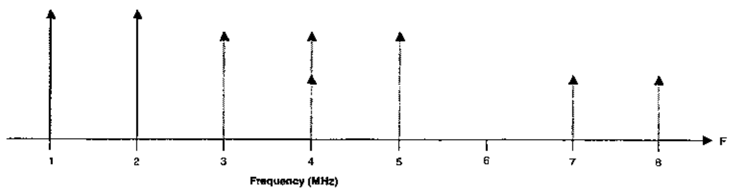


Fig. 1. Spurious mixer products occur when F_1=1MHz and F_2=2MHz are mixed non-linearly.

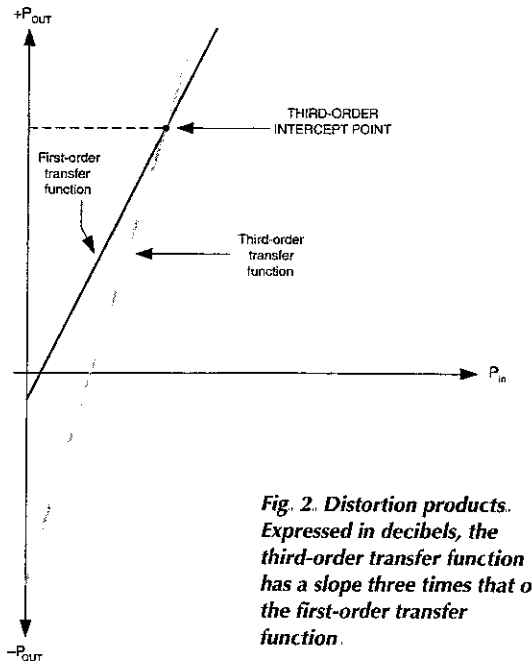


Fig. 2. Distortion products. Expressed in decibels, the third-order transfer function has a slope three times that of the first-order transfer function.

the cube of the input signal level When expressed in decibels e third-order transfer function has a slope three times that the first order transfer function, **Fig. 2**.

Consider the case where two hf signals, F_1=10MHz and F_2=15MHz are mixed together The second-order IPs are 5 and 25MHz; the third-order IPs are 5, 20, 35 and 40MHz; and the fifth-order IPs are 0, 25, 60 and 65MHz. If any of these are inside the passband of the receiver, then they can cause problems.

One such problem is the emergence of 'phantom' signals at the IP frequencies. This effect is seen often when two strong signals F_1 and F_2 exist and can affect the front-end of the receiver, and one of the IPs falls close to a desired signal frequency, F_d If the receiver were tuned to 5MHz, for example, a spurious signal would be found from the F_1-F_2 pair given above.

Another example is seen from strong in-band, adjacent channel signals Consider a case where the receiver is tuned to a station at 9610kHz, and there are very strong signals at 9600kHz and 9605kHz. The near (in-band) IP products are:

Third-order:	9595kHz	(ΔF=15kHz)	
	9610kHz	(ΔF=0kHz)	[on-channel!]
Fifth-order:	9590kHz	(ΔF=20kHz)	
	9615kHz	(ΔF=5kHz)	

Note that one third-order product is on the same frequency as the desired signal, and could easily cause interference if the amplitude is sufficiently high. Other third and fifth-order products may be within the range where interference could occur, especially on receivers with wide bandwidths.

The IP orders are theoretically infinite because there are no bounds on either m or n However, in practical terms each successively higher order IP is reduced in amplitude compared with its next lower order mate. Because of this, only the second-order, third-order and fifth-order products usually assume any importance. Indeed, only the third-order is normally used in receiver specification sheets

Third-order intercept point

It can be claimed that the third-order intercept point, or TOIP, is the single most important specification of a mixer's dynamic performance because it predicts the performance as regards intermodulation, cross-modulation and blocking desensitisation

When a mixer is used in a receiver, the third-order – and higher – intermodulation products are normally very weak They don't exceed the receiver noise floor when the mixer and any preamplifiers are operating in the linear region. But as input signal levels increase, forcing the front-end of the receiver toward the saturated non-linear region, the IP emerges from the noise, **Fig. 3**, and begins to cause problems. When this happens, new spurious signals appear on the band and self-generated interference arises.

Figure 4 shows a plot of the output signal versus fundamental input signal. Note the output compression effect that occurs as the system begins to saturate The dotted gain line continuing above the saturation region shows the theoretical output that would be produced if the gain did not clip.

It is the nature of third-order products in the output signal to emerge from the noise at a certain input level, and increase as the cube of the input level Thus, the slope of the third-order line increases 3dB for every 1dB increase in the response to the fundamental signal.

Although the output response of the third-order line saturates similarly to that of the fundamental signal, the gain line can be continued to a point where it intersects the gain line of the fundamental signal This point is the third-order intercept point.

Note that in Fig. 4, the gain P_O/P_{IN} begins to decrease near the TOIP The measure of this tendency to saturation is called the –1dB compression point, i.e. the point where the gain slope decreases by 1dB

Interestingly enough, one tactic that can help reduce IP levels back down under the noise is the use of an attenuator ahead of the mixer Even a few decibels of input attenuation is often sufficient to drop the IPs back into the noise, while afflicting the desired signals only a small amount

Many modern receivers provide a switchable attenuator ahead of the mixer This practice must be evaluated closely, however, if low-level signals are to be handled. The usual resistive attenuator pad will increase the thermal noise level appearing at the input of the mixer by an amount proportional to its looking back resistance

The IP performance of the mixer selected for a receiver design can profoundly affect the performance of the receiver For example, the second-order intercept point affects the half-IF spur rejection, while the third-order intercept point

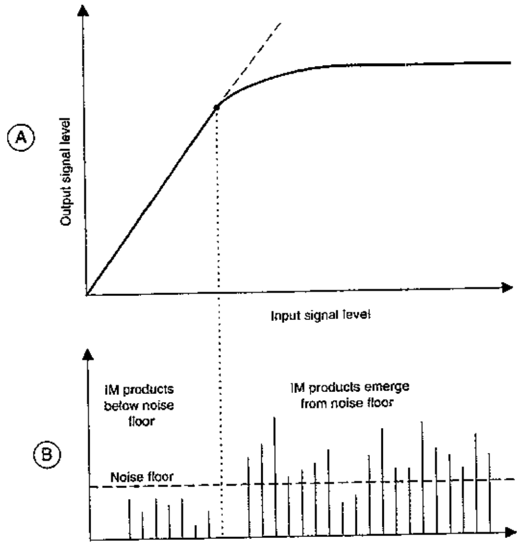


Fig 3 Intermodulation products rise up out of the noise hen critical input level is exceeded.

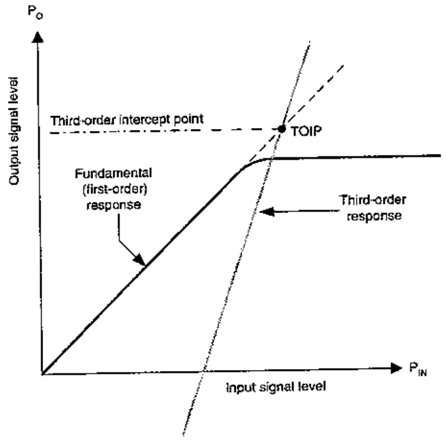


Fig. 4. Third-order intercept point, or TOIP. Note the output compression occurring as the system starts to saturate.

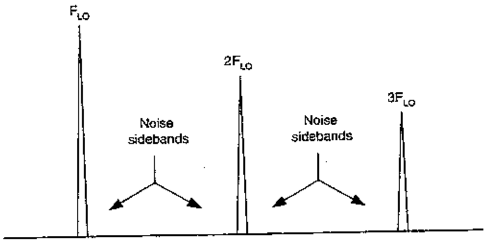
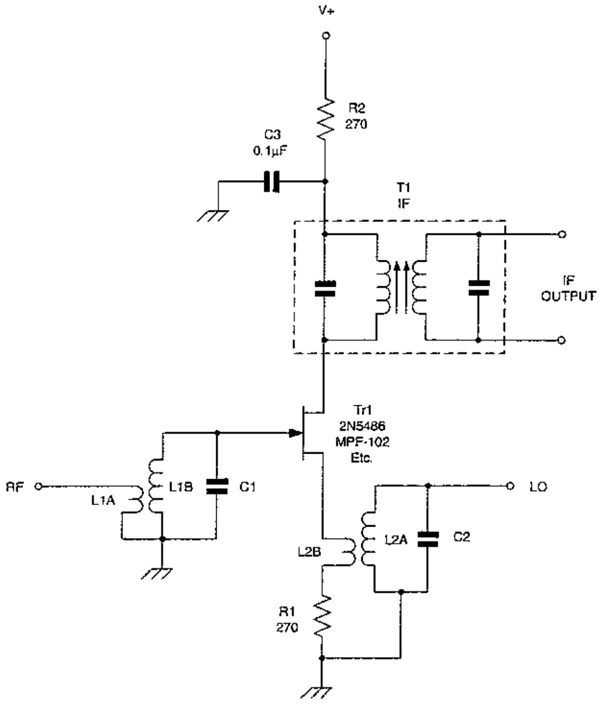


Fig. 5. Noise sidebands surrounding local oscillator and its harmonics can deteriorate sensitivity.

Fig. 6. Single-ended junction fet mixer using a double-tuned LC transformer.



will affect the intermodulation distortion (IMD) performance.

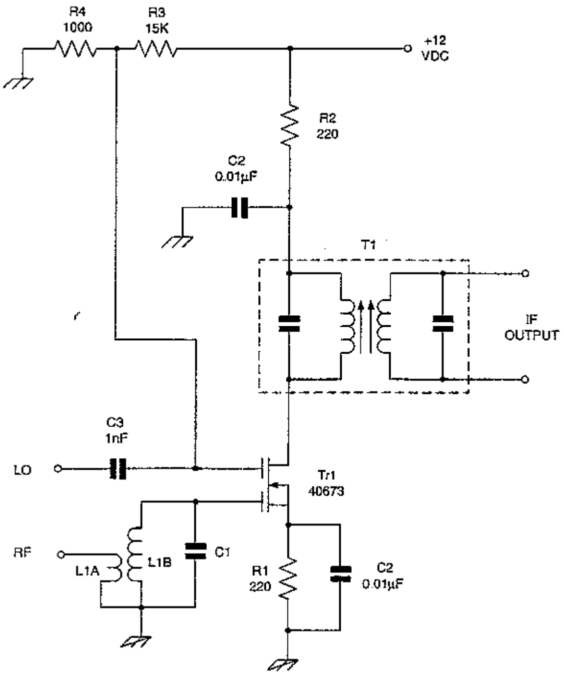
Calculating intercept points

Calculating the *n*th order intercept point can be done using a two-tone test scheme. A test system is created in which two equal amplitude signals, F_A and F_B, are applied simultaneously to the mixer rf input. These signals are set to a standard level of typically –20dBm to –10dBm.

The power of the *n*th intermodulation product, or P_{IMN}, is measured using a spectrum analyser or, if the spectrum analyser is tied up elsewhere, a receiver with a calibrated S-meter. The *n*th intercept point is:

$$IP_N = \frac{NP_A - P_{IMN}}{N-1} \qquad (1)$$

Fig. 7. Single-ended mosfet mixer in which the local-oscillator signal is applied to gate 2.



where IP_N is the intermodulation product of order N, P_A is the input power level, in dBm, of one of the input signals and P_{IMN} is the power level in dBm of the *n*th IM product. Power P_{IMN} is often specified in terms of the receiver's minimum discernible signal specification.

Once the P_A and P_{IM} points are found, any IP can be calculated using the above equation.

Mixer losses

Depending on its design, a mixer may show either loss or gain. The principal loss is conversion loss, which is made up of three elements: mismatch loss, parasitic loss and junction loss, assuming a diode mixer.

Conversion loss is simply a measure of the ratio of the rf input signal level and the signal level appearing at the intermediate-frequency output, P_{IF}/P_{RF}. In some cases, it may be a gain, but for many – perhaps most – mixers there is a loss. Conversion loss L_C is,

$$L_C = L_M + L_P + L_J \qquad (2)$$

where L_C is conversion loss, L_M is the mismatch loss, L_P is the parasitic loss and L_J is the junction loss.

Mismatch loss is a function of the impedance match at the rf and intermediate-frequency ports. Mixer port impedance Z_P and the source impedance Z_S should be matched. If they are not, a voltage-standing-wave ratio, or vswr, will result that is equal to the ratio of the higher impedance to the lower impedance. Depending on which ratio is greater than or equal to 1, $VSWR=Z_P/Z_S$ or $VSWR=Z_S/Z_P$.

The mismatch loss is the sum of rf and intermediate-frequency port mismatch losses. Or expressed in terms of vswr,

$$L_M = 10 \times \left[\log_{10}\left\{\frac{(VSWR_{RF}+1)^2}{4VSWR_{RF}}\right\} + \log_{10}\left\{\frac{(VSWR_{IF}+1)^2}{4VSWR_{IF}}\right\}\right] \qquad (3)$$

Parasitic loss is due to action of the diode's parasitic elements, i.e. series resistance, R_S, and junction capacitance C_J. Junction loss is a function of the diode's *I*-versus-*V* curve. The latter two elements are controlled by careful selection of the diode used for the mixer.

Noise figure

Radio reception is largely an issue of signal-to-noise ratio, also known as SNR. In order to recover and demodulate weak signals the noise figure, or NF, of the receiver is an essential characteristic.

The mixer can be a large contributor to the overall noise performance of the receiver. Indeed, the noise performance of the receiver is seemingly affected far out of proportion to the actual noise performance of the mixer. But a study of signals and noise will show, through Friis' equation, that the noise performance of a receiver or cascade chain of amplifiers is dominated by the first two stages, with the first stage being so much more important than the second stage.

Because of the importance of mixer noise performance, a low noise mixer must be designed or procured. In general, the noise figure of the receiver equipped with a diode mixer first stage – as is common in microwave receivers – is,

$$NF = L_C + IF_{NF} \qquad (4)$$

where NF is the overall noise figure, L_C is the conversion loss and IF_{NF} is the noise figure of the first IF amplifier stage.

To obtain the best overall performance from the perspective of the mixer, the following should be observed:

1. Select a mixer diode with a low noise figure. This will address the junction and parasitic losses.
2. Ensure the impedance match of all mixer ports.
3. Adjust the local-oscillator power level for minimum con-

version loss. Local oscillator power is typically higher than maximum rf power level

Noise balance

There is noise associated with the local-oscillator signal, and that noise can be transferred to the IF in the mixing process. The tendency of the mixer to transfer AM noise to the IF is called its noise balance.

In some cases, this transferred noise results in loss that is more profound than the simple conversion loss, so should be evaluated when selecting a mixer.

The total noise picture, **Fig. 5**, includes not simply the AM noise sidebands around the local oscillator frequency, but also the noise sidebands around the local oscillator harmonics The latter can be eliminated by imposing a filter between the local-oscillator output and the mixer's local-oscillator input.

The noise sidebands around the local oscillator itself, however, are not easily suppressed by filtering because they are close in frequency to F_{LO}. The use of a balanced mixer, however, can suppress all of the local-oscillator signal in the output, and that includes the noise sidebands. In the usual way noise balance is specified, the higher the number in decibels the more suppression of local-oscillator AM noise.

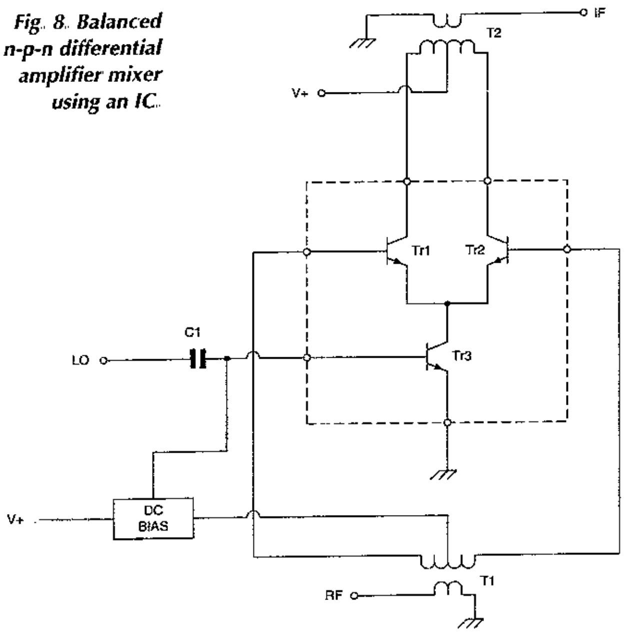


Fig. 8. Balanced n-p-n differential amplifier mixer using an IC.

Single-ended active mixer circuits

So far, the only mixer circuit that has been explicitly discussed is the diode mixer The diode is a general category called a switching mixer because the local-oscillator switches the diode in and out of conduction.

Now I will look at active single-ended unbalanced mixers. **Figure 6** shows the circuit of a simple single-ended unbalanced mixer based on a junction field effect transistor, or JFET, such as the *MPF-102* or *2N5486*.

The rf signal is applied to the gate, while the local-oscillator signal is applied to the source. If the local-oscillator signal has sufficient amplitude to cause non-linear action, then it will permit the JFET to perform as a mixer.

Note that both the rf and local oscillator ports are fitted with bandpass filters to limit the frequencies that can be applied to the mixer. Because these mixers tend to have rather poor local-oscillator-to-rf and rf-to-local-oscillator isolation, these tuned filters help improve the port isolation by preventing the local-oscillator from appearing in the rf output and the rf from being fed to the output of the local-oscillator source

In many practical cases, the local oscillator filter may be eliminated because it is difficult to make a filter that will track a variable local-oscillator frequency In some cases, the receiver designer will use an untuned bandpass filter, while in others the output of the local-oscillator is applied directly to the source of the JFET through either a coupling capacitor or an untuned rf transformer.

The output of the unbalanced mixer contains the full spectrum of $mF_{RF}\pm nF_{LO}$ products, so a tuned filter is needed here also The drain terminal of the JFET is the IF port in this circuit

The usual case is to use either a double-tuned *LC* transformer, T_1 in Fig 6, or some other sort of filter. Typical non-*LC* filters used in receivers include ceramic and quartz crystal filters, and mechanical filters

A MOSFET version of the same type of circuit is shown in **Fig. 7** In this circuit, a dual-gate MOSFET such as a *40673* is the active element The rf is applied to gate 1, and the

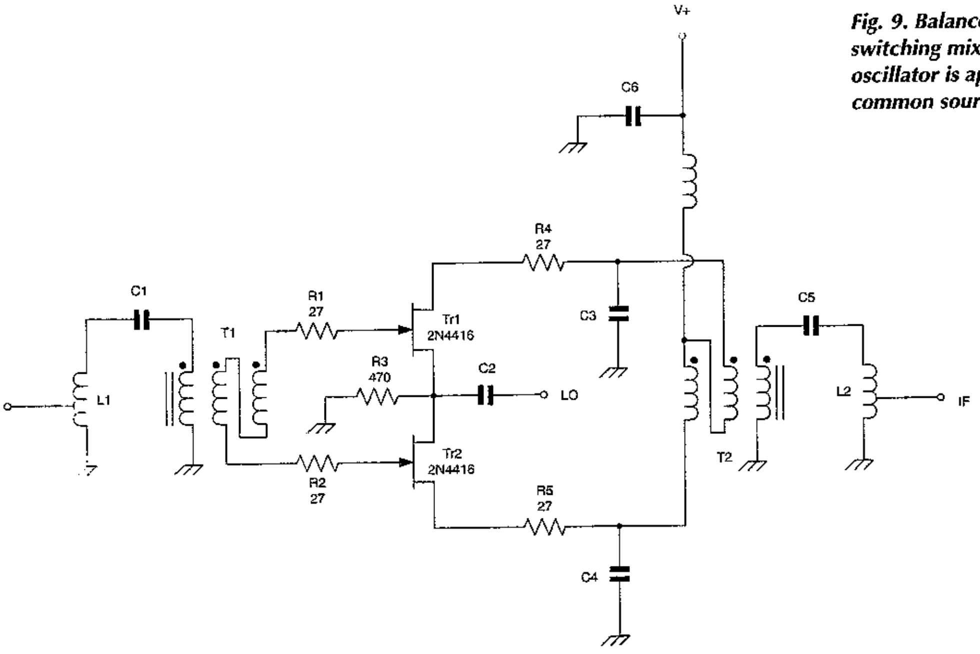


Fig. 9. Balanced junction fet switching mixer. Here the local oscillator is applied to the common source.

local-oscillator is applied to gate 2.

Local oscillator signal level needs to be sufficient to drive Tr_1 into non-linear operation. A resistive voltage divider R_3/R_4 is used to provide a dc bias level to gate 2. The source terminal is bypassed to ground for rf, and is the common terminal for the mixer.

In this particular case the local-oscillator input is broadband, and is coupled to the local-oscillator source through capacitor C_3. A resonant bandpass filter, L_{1B}/C_1, on the other hand, tunes the rf input.

Balanced active mixers

There is a number of balanced active mixers that can be selected. Many of these forms are now available in integrated circuit form

Because of the intense activity being seen in the development of telecommunications equipment – cellular, PCS and other types – there is a lot of IC development being done in this arena.

One of the earliest types of rf IC on the market was the differential amplifier. **Figure 8** shows the use of one of these ICs are a mixer stage. Two transistors, $Tr_{1,2}$, are differentially connected by having their emitter terminals connected together to a common current source, Tr_3.

The rf signal is applied to the bases of Tr_1 and Tr_2 differentially through transformer T_1. The local-oscillator signal is used to drive the base terminal of the current source transistor, Tr_3. The collectors of Tr_1 and Tr_2 are differentially connected through a second transformer, T_2, which forms the IF port

Figure 9 shows one rendition of the Sabin double-balanced mixer used in hf receivers. It offers a noise figure of about 3dB. Dating from around 1970, the Sabin mixer features a push-pull pair of high pinch-off voltage junction fets, $Tr_{1,2}$. These are connected in a common source configuration. The local-oscillator signal is applied to the common source in a manner similar to Fig. 7.

The gate circuits of $Tr_{1,2}$ are driven from a balanced transformer, T_1. This transformer is trifilar wound, usually on either a toroid or binocular balun core. The dots on the transformer windings indicate the phase sense of the winding

Note that the gate of Tr_1 is fed from a dotted winding end, while that of Tr_2 is fed from a non-dotted end. This arrangement ensures that the signals will be 180° out of phase, resulting in the required push-pull action. Some input filtering and impedance matching the 1.5kΩ junction fet input impedance to a 50 or 75Ω system impedance, as required, is provided by L_1/C_1.

The IF output is similar to the rf input. A second trifilar transformer T_2 is connected such that one drain is to a dotted winding end and other is to a non-dotted end of T_2

Compare the sense of the windings of T_1 and T_2 in order to avoid signal cancellation due to phasing problems. Intermediate-frequency filtering and impedance matching is provided by C_5 and L_2. The tap on L_2 is adjusted to match the 5.5kΩ impedance of the junction fets to system impedance.

In the final article of this set I will take a look at a number of circuits for single and double balanced mixers, including the passive diode version that is so popular with designers ■