

REPORT ITU-R SM.2021

**PRODUCTION AND MITIGATION OF INTERMODULATION
PRODUCTS IN THE TRANSMITTER**

(Question ITU-R 211/1)

(2000)

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1 Introduction

There are various types of intermodulation that can be found. In radio systems, these are manifested in a number of ways and defined as the following five types in Rec. ITU-R SM.1446:

Type 1: Single channel intermodulation: where the wanted signal is distorted by virtue of non-linearities in the transmitter.

Type 2: Multichannel intermodulation: where the wanted signals of multi channels are distorted by virtue of non-linearities in the same transmitter.

Type 3: Inter transmitter intermodulation: where one or more transmitters on a site intermodulate, either within the transmitters themselves or within a non-linear component on site to produce intermodulation products.

Type 4: Intermodulation due to active antennas: the multicarrier operating mode of an active antenna, along with the non-linearity of amplifiers, originates spurious emissions under the form of intermodulation signals.

Type 5: Intermodulation due to passive circuits: where transmitters share the same radiating element and intermodulation occurs due to non-linearities of passive circuits.

The generation and mitigation of these intermodulation products are described in the following sections in more detail. Some examples of intermodulation products generated at radio sites are given. Measurement techniques are referred in Rec. ITU-R SM.1446. A comprehensive list of useful literature is attached at the end of the Report including references for the measurement of intermodulation Types 1 to 3 [ETSI, 1997; Shahid *et al.*, 1996; Bhargava *et al.*, 1981; ITU-R Handbook on satellite communications – fixed-satellite service (Appendix 2-1, § 5); Heathman, 1989; Bond et Meyer, 1970; Shimbo, 1971; Saleh, 1982; Wassermann *et al.*, 1983; Tondryk, 1991; Kaeadar, 1986; IEES, 1996; ETSI, 1995].

Instead of intermodulation products the expression intermodulation noise is also used in order to reflect digital modulation formats.

2 Generation of intermodulation

Intermodulation has classically been a major determinant of transmitter performance for amplitude modulated services, such as single sideband (SSB) or independent sideband (ISB). Theoretically, it does not apply to any constant envelope transmission, although in practice, practical implementation limitations lead to some of such modulation techniques not providing absolutely constant envelope modulation, and thus requiring linear amplification if spectral regrowth is to be avoided.

2.1 Intermodulation products due to discrete frequencies

The following approach [Chadwick, 1986] is classical and a complete analysis for input signal which can be represented by discrete frequencies like all analogue signals in the time domain. It may be also helpful for the basic understanding of the generation of intermodulation products.

An amplifier can be characterized by a Taylor series of the generalized transfer function [Chadwick, 1986]

$$i_0 + k_1 e_{IN} + k_2 e_{IN}^2 + k_3 e_{IN}^3 + k_4 e_{IN}^4 + k_5 e_{IN}^5 + \dots$$

where i_0 is the quiescent output current, k_1 , k_2 , etc. are coefficients and e_{IN} represents the input signal. When two sinusoidal frequencies $\omega_1 = 2\pi f_1$ and $\omega_2 = 2\pi f_2$ of the amplitude a_1 and a_2 are applied to the input of the amplifier, the input signal is:

$$e_{IN} = a_1 \cos \omega_1 t + a_2 \cos \omega_2 t$$

and the output i_{OUT} may be shown to be the sum of the DC components:

$$i_{OUT} = i_0 + \frac{k_2}{2} (a_1^2 + a_2^2) + \frac{k_4}{8} (3a_1^4 + 12a_1^2 a_2^2 + 3a_2^4)$$

fundamental components,:

$$+ \left(k_1 a_1 + \frac{3}{4} k_3 a_1^3 + \frac{3}{2} k_3 a_1 a_2^2 + \frac{5}{8} k_5 a_1^5 + \frac{15}{4} k_5 a_1^3 a_2^2 + \frac{15}{8} k_5 a_1 a_2^4 \right) \cos \omega_1 t$$

$$+ \left(k_1 a_2 + \frac{3}{4} k_3 a_2^3 + \frac{3}{2} k_3 a_1^2 a_2 + \frac{5}{8} k_5 a_2^5 + \frac{15}{4} k_5 a_1^2 a_2^3 + \frac{15}{8} k_5 a_1^4 a_2 \right) \cos \omega_2 t$$

2nd order components:

$$+ \left(\frac{1}{2} k_2 a_1^2 + \frac{1}{2} k_3 a_1^4 + \frac{3}{2} k_4 a_1^2 a_2^2 \right) \cos 2\omega_1 t$$

$$+ \left(\frac{1}{2} k_2 a_2^2 + \frac{1}{2} k_3 a_2^4 + \frac{3}{2} k_4 a_1^2 a_2^2 \right) \cos 2\omega_2 t$$

$$+ \left(k_2 a_1 a_2 + \frac{3}{2} k_4 a_1^3 a_2 + \frac{3}{2} k_4 a_1 a_2^3 \right) \cos (\omega_1 \pm \omega_2) t$$

3rd order components:

$$+ \left(\frac{1}{4} k_3 a_1^3 + \frac{5}{16} k_5 a_1^5 + \frac{5}{4} k_5 a_1^3 a_2^2 \right) \cos 3\omega_1 t$$

$$+ \left(\frac{1}{4} k_3 a_2^3 + \frac{5}{16} k_5 a_2^5 + \frac{5}{4} k_5 a_1^2 a_2^3 \right) \cos 3\omega_2 t$$

$$+ \left(\frac{3}{4} k_3 a_1^2 a_2 + \frac{5}{4} k_5 a_1^4 a_2 + \frac{15}{8} k_5 a_1^2 a_2^3 \right) \cos (\omega_1 \pm 2\omega_2) t$$

$$+ \left(\frac{3}{4} k_3 a_1 a_2^2 + \frac{5}{4} k_5 a_1 a_2^4 + \frac{15}{8} k_5 a_1^3 a_2^2 \right) \cos (\omega_2 \pm 2\omega_1) t$$

4th order components:

$$+ \frac{1}{8} k_4 a_1^4 \cos 4\omega_1 t + \frac{1}{8} k_4 a_2^4 \cos 4\omega_2 t$$

$$+ \frac{1}{2} k_4 a_1^3 a_2 \cos (3\omega_1 \pm \omega_2) t + \frac{3}{4} k_4 a_1^2 a_2^2 \cos (2\omega_1 \pm 2\omega_2) t + \frac{1}{2} k_4 a_1 a_2^3 \cos (\omega_1 \pm 3\omega_2) t$$

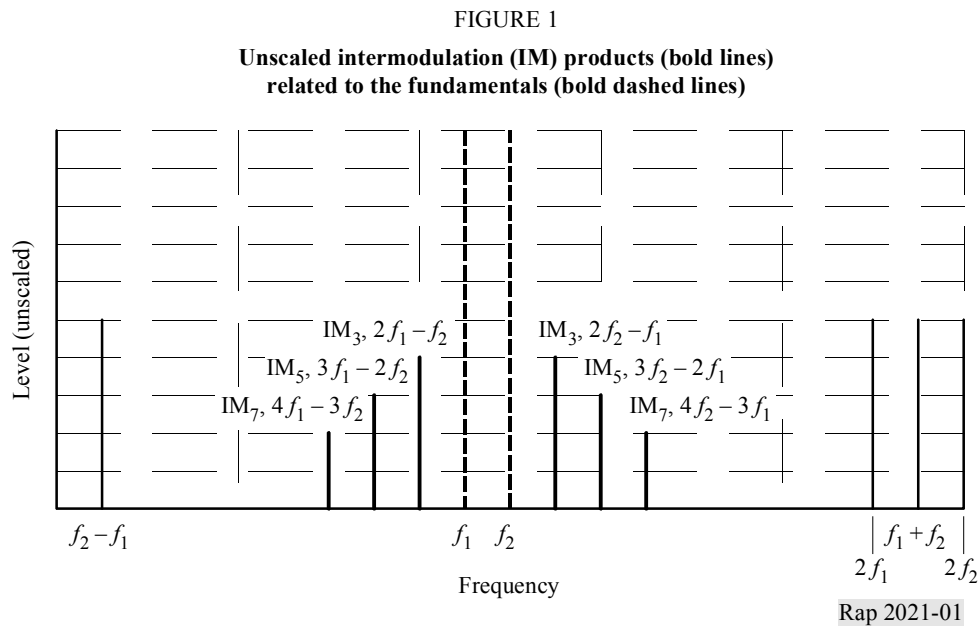
and 5th order components:

$$+ \frac{1}{16} k_5 a_1^5 \cos 5\omega_1 t + \frac{1}{16} k_5 a_2^5 \cos 5\omega_2 t$$

$$+ \frac{5}{16} k_5 a_1^4 a_2 \cos (4\omega_1 \pm \omega_2) t + \frac{5}{8} k_5 a_1^3 a_2^2 \cos (3\omega_1 \pm 2\omega_2) t$$

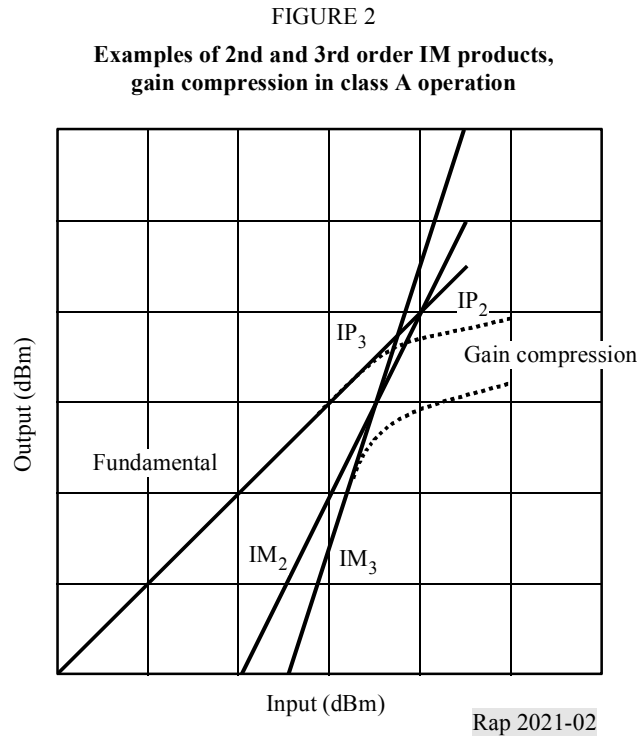
$$+ \frac{5}{8} k_5 a_1^2 a_2^3 \cos (2\omega_1 \pm 3\omega_2) t + \frac{5}{16} k_5 a_1 a_2^4 \cos (\omega_1 \pm 4\omega_2) t$$

The series may be expanded further for terms in $k_6 e_{IN}^6$ etc. if desired. The relationships between the different products are shown in Fig. 1. It can be seen from this Figure and the equations that all the even order terms produce outputs at harmonics of the input signal and that the sum and difference products are well removed in frequency far from the input signal. The odd order products, however, produce signals near the input frequencies $f_1 \pm 2f_2$ et $f_2 \pm 2f_1$. Therefore, the odd order intermodulation products cannot be removed by filtering, only by improvement in linearity.



Assuming class A operation, $a_1 = a_2$ and k_4, k_5 are very small. The 3rd order intermodulation product IM_3 becomes proportional to a^3 . That means that the cube of the input amplitude and the graph of the intermodulation products will have a slope of 3 in logarithmic scale while the wanted signal will have the slope of 1 (see Fig. 2). Second order products IM_2 can be similarly calculated, and the graph for these has a slope of two. The points where these graphs cross are called 3rd order intercept point IP_3 and 2nd order intercept point IP_2 , respectively. IP_3 is the point where the intermodulation product is equal to the fundamental signal. This is a purely theoretical consideration, but gives a very convenient method of comparing devices. For example, a device with intermodulation products of -40 dBm at 0 dBm input power is to be compared with one having intermodulation products of -70 dBm for -10 dBm input. By reference to the intercept point, it can be seen that the two devices are equal.

As the level of the input signal increases, a point is eventually reached at which the output cannot increase, dB for dB, with the input. This is gain compression, and is important in defining the dynamic range of the device. For example, assuming an amplifier with 20 dBm intercept point and at 0 dBm input to obtain a 40 dB intermodulation ratio, but because the devices' input/output characteristics are not linear at this input level, the expected intermodulation ratio is not obtained. If, however, the compression point is a few dB higher, then the intermodulation ratio of 40 dB could be obtained. In the case of class AB operation different characteristics may occur as plotted in Fig. 2, especially at lower input signals.



2.2 Intermodulation noise due to continuous frequency spectrum

The classical description of intermodulation at analogue radio systems deals with a two-frequency input model to a memoryless non-linear device. This non-linear characteristic can be described by a function $f(x)$, which yields the input-output relation of the element device. The function, f , is usually expanded in a Taylor-series and thus produces the harmonics and as well the linear combinations of the input frequencies. This classical model is well suited to analogue modulation schemes with dedicated frequency lines at the carrier frequencies. The system performance of analogue systems is usually measured in terms of signal-to-noise (S/N) ratio, and the distorting intermodulation signal can adequately be described by a reduction of S/N .

With digital modulation methods the situation is changed completely. Most digital modulation schemes have a continuous signal spectrum without preferred lines at the carrier frequencies. The system degradation due to intermodulation is measured in terms of bit error ratio (BER) and depends on a variety of system parameters, e.g. the special modulation scheme which is employed.

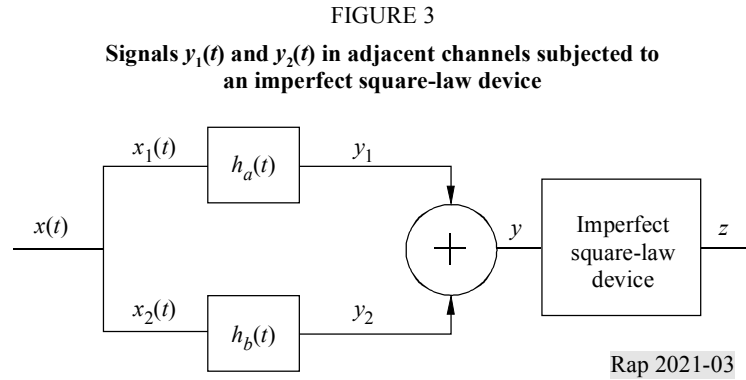
For estimation of the system performance in terms of BER a rigorous analysis of non-linear systems is required. There are two classical methods for the analysis and synthesis of non-linear systems: the first one carries out the expansion of the signal in a Volterra series [Schetzen, 1980]. The second due to Wiener uses special base functionals for the expansion.

Both methods lead to a description of the non-linear system by higher order transfer functions having n input variables depending on the order of the non-linearity. A more detailed description and two examples are given in Annex 1.

The block diagram of example 1 is shown in Fig. 3. The two data signals $x_1(t)$ and $x_2(t)$ are linearly filtered by the devices with the impulse responses $h_a(t)$ and $h_b(t)$ in adjacent frequency bands. The composite summed signal y is hereafter distorted by an imperfect square-law device which might model a transmit-amplifier. The input-output relation of the non-linear device is given by:

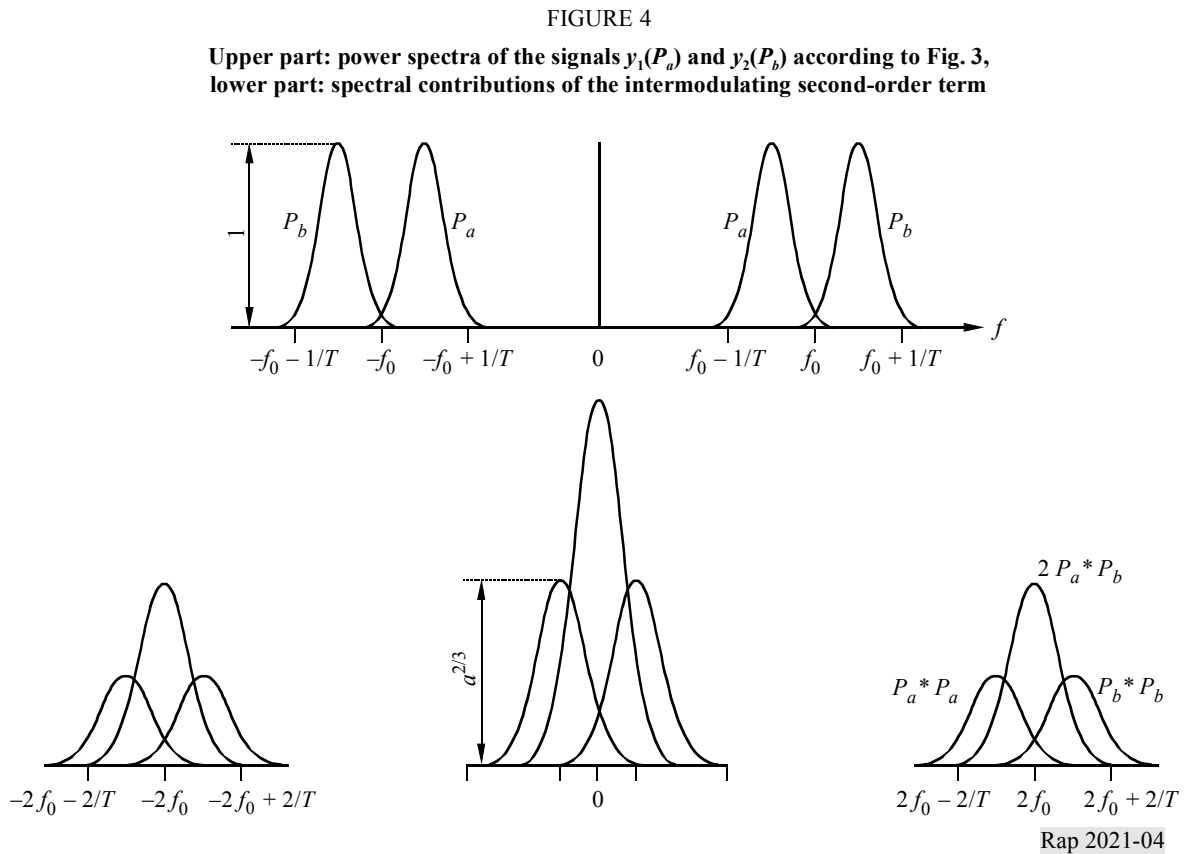
$$z(t) = y(t) + ay^2(t)$$

The input signals $x_1(t)$ and $x_2(t)$ are originated from a single signal $x(t)$, because of the spectral separation by the filters $h_a(t)$ and $h_b(t)$.

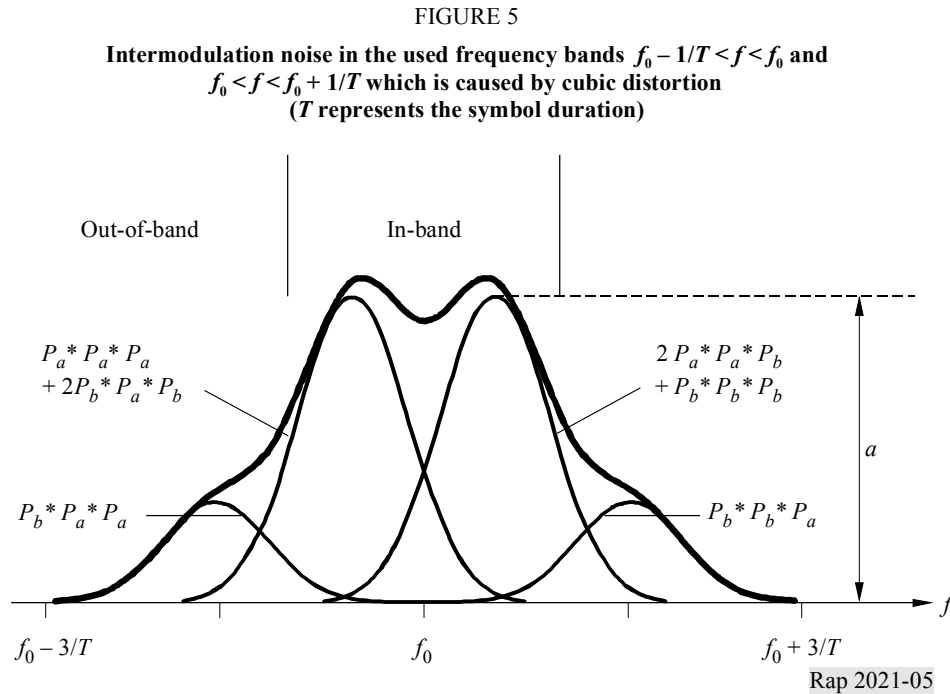


The output signal $z(t)$ including the intermodulation noise is plotted in Fig. 4. For RF-modulated signals the intermodulation distortion in the proper frequency band is caused by non-linearities of third order. For this reason the imperfect square-law device in Fig. 3 is now replaced by an imperfect cubic device with the input-output relation:

$$z(t) = y(t) + ay^3(t)$$



there are several contributions of the intermodulation noise falling into the used channels near f_0 . The different parts $P_a * P_a * P_a \dots P_b * P_b * P_b$ are plotted in Fig. 5. The thick line shows the sum of the distortions.



3 Mitigation techniques

A number of techniques have been developed to reduce intermodulation in transmitter power amplifiers, and some of these are briefly described. However, they are not considered as an exhaustive list.

In certain standards [ETSI, 1994 and 1998], distinction is often made between active intermodulation due to non-linear components in the transmitters themselves and passive intermodulation. Passive intermodulation is caused, e.g., by metallic contacts in masts. Antenna hardware can be a source of problems when high field strengths occur on the site and cause them to radiate intermodulation products which disturb equipment at the same site, or at a neighbouring site. No standards exist that are able to specify reasonable limits for such effects. The intermodulation factors used to make these calculations are affected by many parameters and they also depend on electrical resonances in the components which make up the mast and antenna arrays. These products can be radiated from the site.

The overall loss, A_{CI} , between a transmitter providing the unwanted emissions giving rise to the intermodulation product is given by the sum of:

$$A_{CI} = A_C + A_I$$

where A_C is the coupling loss defined as the ratio of the power emitted from one transmitter to the power level of that emission at the output of another transmitter which may produce the unwanted intermodulation product. The intermodulation conversion loss, A_I , is the ratio of power levels of the interfering signal from an external source and the intermodulation product, both measured at the output of the transmitter (ITU-R Report M.739 – Interference due to intermodulation products in the land mobile service between 25 and 1 000 MHz).

Using this definition, mitigation of intermodulation products means increasing the overall loss A_{CI} . It is obvious that a reduction of the non-linearity, particularly of the odd-numbered orders, will improve the overall performance and increase the value of intermodulation conversion loss. Such techniques considered in § 3.1 can be used to reduce the intermodulation products of the Types 1 and 2 and can be implemented in the transmitters themselves. Possibilities for increasing the coupling loss, e.g. by increasing the spatial separation, are described in § 3.2. These mitigation measures which are known as radio site shielding engineering are applicable for intermodulation Type 3. In contrast to § 3.1 these measures are not used in the transmitters themselves.

Other mitigation measures are briefly quoted in § 3.3. Some examples of intermodulation Type 3 at the transmitter site are depicted in § 3.4 for illustration.

3.1 Suppression at transmitters

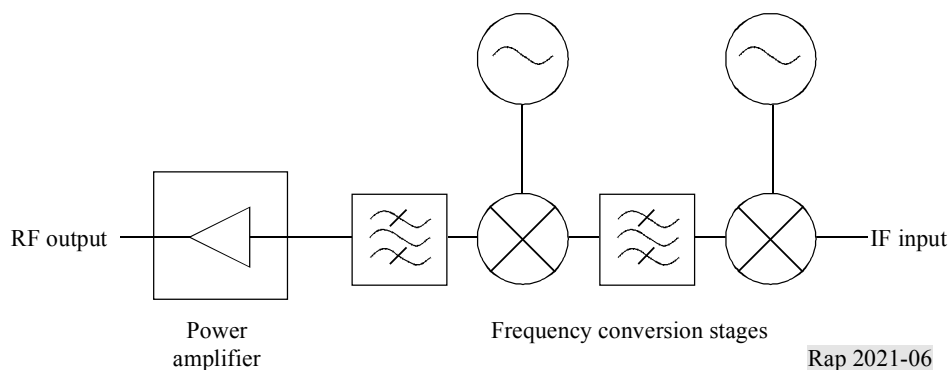
These intermodulation products are part of the unwanted emissions as defined in No. S1.145 of the Radio Regulations. In the following a typical transmitter architecture is introduced before discussing mitigation techniques.

3.1.1 Transmitter architecture

The RF architecture of radio transmitters often takes the form shown in the simplified block diagram of Fig. 6. The modulated input signal is generated at an IF, then frequency translated in one or more mixing and filtering stages to the final frequency.

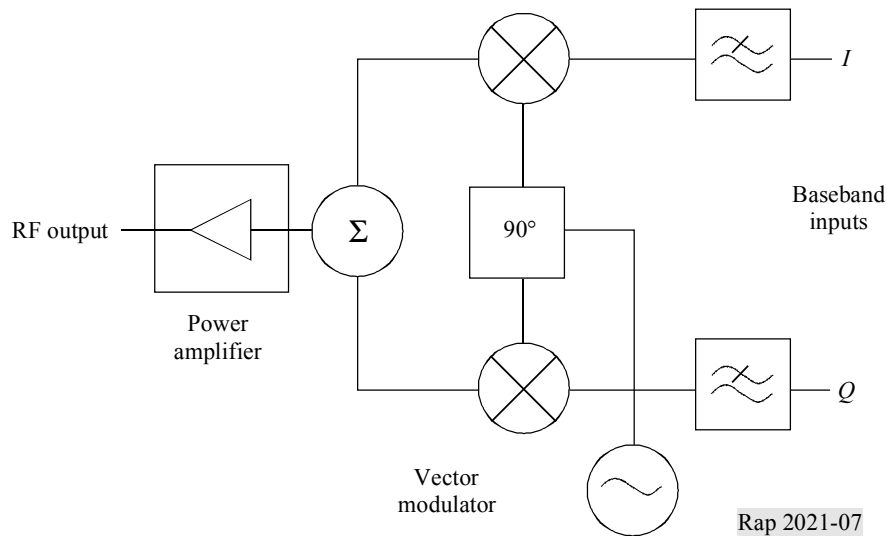
A common problem with this arrangement is that each mixing process will produce many spurious products, as well as the main sum and difference frequency components. These arise through mixing of the local oscillator (LO) harmonics with harmonics of the IF input, often referred to as $m \times n$ products. Although the LO harmonics are unavoidable due to the modulating action of the mixer LO port, the IF harmonics can be reduced by ensuring that the IF port is operated well below compression. However, in practice, a compromise must be reached between linearity and S/N , so the spuri can never be completely eliminated. Spurious products which fall at offsets far removed from the wanted frequency can be suppressed through filtering, but those close to the carrier will not be attenuated.

FIGURE 6
Typical up-conversion transmitter architecture



One way of mitigating this problem is to generate the wanted signal directly at the final frequency using a vector modulator, as shown in Fig. 7. In this case, in-phase and quadrature (I and Q) baseband signals are used to directly modulate a carrier at the output frequency. Although spectral spreading of the signal into the adjacent channels can still occur, the harmonic mixing effect is eliminated, since there is only a single carrier component applied to the mixers.

FIGURE 7
Vector modulator transmitter architecture



A drawback with this arrangement is that there will be a finite carrier leakage to the output, typically suppressed by about 30 dB relative to the wanted signal. Usually this is of no consequence, but in cases where better carrier suppression is required, it is necessary to adjust the d.c. bias on the *I* and *Q* inputs to null the carrier.

While the arrangement illustrated in Fig. 7 utilizes two bi-phase AM modulators, it is equally possible to use four uniphase modulators, and four orthogonal channels.

A more complex, but more flexible, approach is to use a single channel incorporating a digitally controlled attenuator, and a digitally controlled phase-shifter. These two components are driven by the baseband input by way of a look-up table, allowing the direct generation of virtually any (digital) modulation scheme. It might be noted that the carrier-frequency amplitude/phase shifter is the same component required for use in active antenna beam-forming arrays.

3.1.2 Filtering

Filtering (generally bandpass filtering) of the transmitter output can be used in conjunction with the other techniques discussed in this Report to reduce the residual spurious output levels. The choice of the type of filter to be used is, as usual, a compromise between a number of interacting, usually conflicting, requirements such as out-of-band rejection, bandpass attenuation, time domain response, size, weight, cost, etc.

Filter designs are usually based on the classical analytically derived categories such as Butterworth, Chebyshev, etc. Some of these categories are optimized for one of its characteristics at the expense of others, and some provide compromises between characteristics as in Table 1:

TABLE 1

Category	Optimized parameter	Sacrificed parameter
Butterworth	Bandpass amplitude flatness	Out-of-band rejection
Chebyshev	Out-of-band rejection	Bandpass amplitude flatness and attenuation
Bessel	Bandpass delay flatness	Out-of-band rejection
Elliptic (Cauer)	Close-in out-of-band rejection (theoretically infinite at spot frequencies)	Out-of-band rejection away from spot frequencies