Microwave Filter Design from a Systems Perspective

Ian Hunter, Richard Ranson, Andrew Guyette, and Alaa Abunjaileh

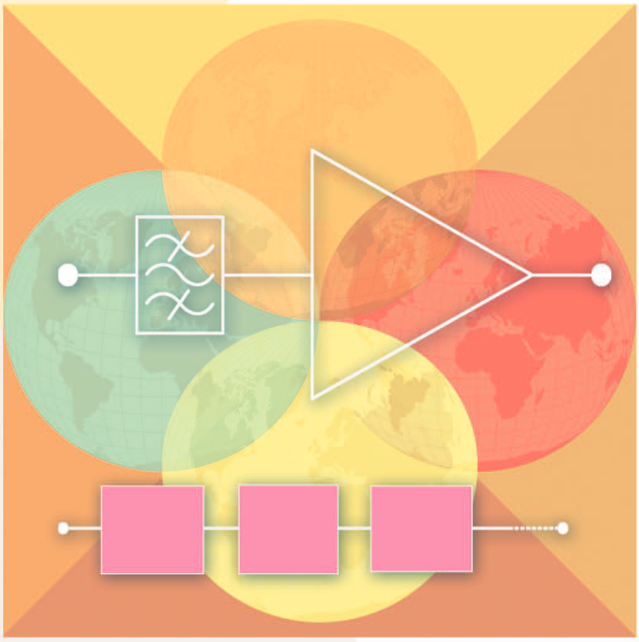
The subject of microwave filter design is a huge one, and it is not possible—or desirable—to give a complete summary here. Details are given in review papers [1] and [2] and books [3] and [4]. The intention of this article is merely to provide the reader with some basic concepts that will enable them to better make intelligent decisions when specifying filters.

We will restrict our discussion to the most common type of filter, the bandpass filter. Bandpass filters are intended to minimize the signal attenuation in one band of frequencies (the passband), while achieving a specified attenuation in another band (the stopband). They are typically designed by applying a frequency transformation to a low-pass prototype network (LPP). The LPP is a ladder-type filter which is normalized to have a low-pass response cutting off a $= 1$ rad/s and operating in a $1 - \Omega$ system impedance. This is shown in Figure 1.

This type of "all-pole" network may be designed to have a variety of transfer functions for specified amplitude or phase characteristics. However, by far the most common transfer function is the Chebyshev (or the modifications), where

$$|S_{12}|^2 = \frac{1}{1 \quad \varepsilon^2 T_N^2(\ )},$$

in which $T_N(\ ) = \cos[N \cos^{-1}(\ )]$.



In fact, $T_N(\ )$ is a Chebyshev polynomial of degree N, and the transfer function is shown in Figure 2.

The Chebyshev transfer function is widely used because it has a relatively high selectivity, i.e., rate of change of response from passband to stopband. It is also relatively insensitive to element value tolerances. Note that this is a simplification; in reality, most

Ian Hunter (I.C.Hunter@leeds.ac.uk), Richard Ranson, Andrew Guyette, and Alaa Abunjaileh are with the Institute of Microwaves and Photonics, School of Electronic and Electrical Engineering, of the University of Leeds, UK.

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modern filters are designed for a generalized Chebyshev response, which enables group-delay equalization, finite frequency attenuation poles, and asymmetric frequency responses. The degree N of the Chebyshev filter determines its selectivity. Consider the bandpass filter specification shown in Figure 3.

Here the passband is from $_1$ to $_2$ and the center of the passband is at $_0$. We will specify a stopband attenuation of LA dB at $_3$ and a similar value at the (symmetrically located) $_4$.

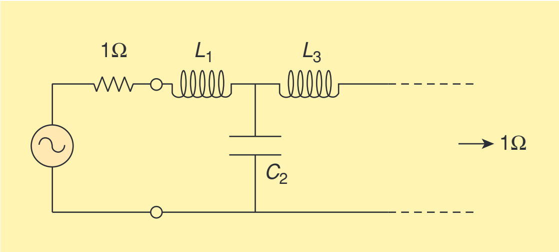

Figure 1. *Low pass prototype network.*

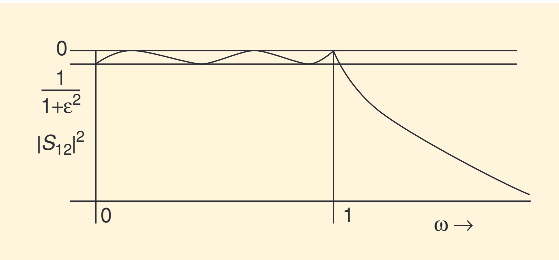

Figure 2. *Chebychev LPP transfer function.*

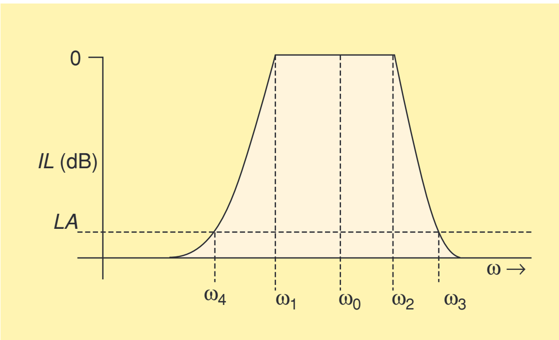

Figure 3. *Bandpass filter specification.*

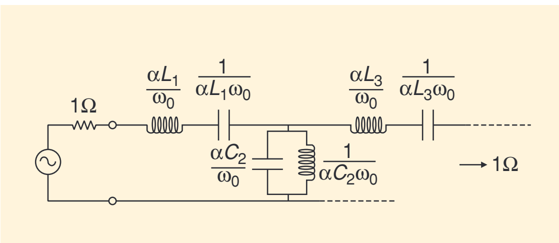

Figure 4. *Bandpass transformation of LPP.*

Define a selectivity factor S, where

$$S = \frac{\ _3 - \ _0}{\ _2 - \ _0}.$$

Then

$$N \geq \frac{L_A + L_R + 6}{20Log_{10}[S + \sqrt{S^2 - 1}]},$$

where L_R is the passband return loss specification.

A closely related parameter is the shape factor, which is the ratio of the width of an attenuation level close to the passband loss to another attenuation level width close to the maximum stopband requirement. Typical evaluation points are at 6 and 60 dB, where a good filter would have a shape factor of close to two compared to the ideal value of unity.

As an example, consider a bandpass filter with a center frequency of 10 GHz and a passband bandwidth of 100 MHz. A stopband attenuation of 60 dB is required at 9.9 and 10.1 GHz. A minimum passband return loss of 20 dB is specified.

$$S = \frac{10.1 - 10}{10.05 - 10} = 2.$$

$LA = 60,\ LR = 20,$ giving

$$N \geq 7.518.$$

Thus, an eight-pole Chebyshev filter is required.

Having decided on the degree of the Chebyshev LPP, we must now convert the LPP to a bandpass filter. For a lumped-element realization, this may be achieved by applying the transformation

$$\alpha\left[\frac{}{\ _0} - \frac{\ _0}{}\right],$$

where $_0$ is the angular center frequency of the filter and

$$\alpha = \frac{\ _0}{\Delta},$$

in which Δ is the passband bandwidth.

This transformation has the effect of mapping frequencies $= \pm 1$ in the LPP to $_2$ and $_1$ in the bandpass filter. Furthermore, it converts inductors in the LPP in Figure 1 to series resonant circuits and capacitors to parallel resonant circuits, as shown in Figure 4.

Thus, the bandpass filter consists of a cascade of series and shunt resonant circuits. However, this is a lossless network and we know that, in reality, resonators always have a finite amount of dissipation loss. This is normally described by the unloaded quality factor (or Q factor) of resonators, denoted by Q_u.

For a series resonant circuit, we can associate resistance with the inductor, giving an equivalent circuit shown in Figure 5(a).

The Q factor of this circuit at resonance is

$$Q_u = \frac{\omega_0 L}{R},$$

or

$$R = \frac{\omega_0 L}{Q_u}.$$

Similarly, the lossy parallel resonant circuit shown in Figure 5(b) has a Q factor at resonance given by

$$Q_u = \frac{\omega_0 C}{G},$$

or

$$G = \frac{\omega_0 C}{Q_u}.$$

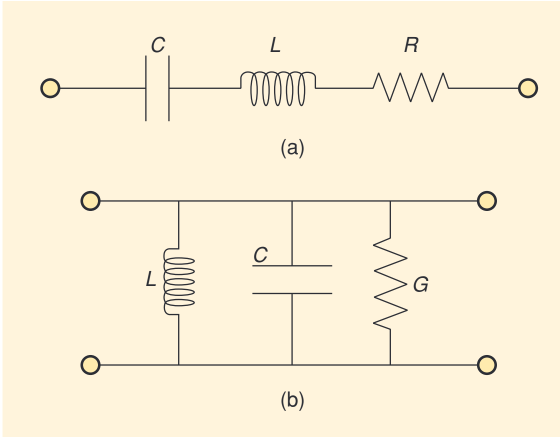

Figure 5. *(a) Lossy series LC circuit. (b) Lossy parallel resonant circuit.*

Now, if the bandpass filter shown in Figure 4 has finite Q_u, then we can replace its equivalent circuit with one containing lossy resonators. Furthermore, at the center frequency, ω_0, all the series resonators are short circuits and all the parallel resonators are open circuits. We are simply left with the resistive circuit shown in Figure 6.

Here,

$$R_1 = \frac{\omega_0}{Q}\left(\frac{\alpha L_1}{\omega_0}\right) = \frac{\alpha L_1}{Q_u},$$

$$G_2 = \frac{\alpha L_3}{Q_u},$$

$$R_3 = \frac{\alpha L_3}{Q_u}.$$

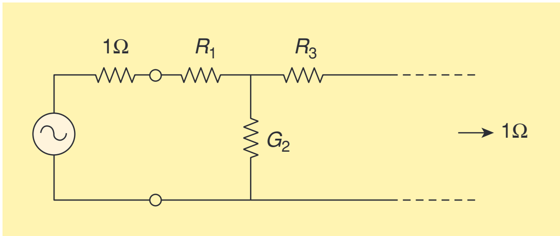

Figure 6. *Equivalent circuit of a lossy bandpass filter at* ω_0.

Thus, at the center frequency ω_0, the filter looks like a resistive attenuator. We can see that the losses are proportional to α and inversely proportional to the unloaded resonator Q factor.

In general, the midband insertion loss of a bandpass filter is given by

$$LA_0 = \frac{4.343\, f_0}{\Delta_f Q_u} \sum_{R=1}^{N} g_r,$$

where the summation is the sum of the element values in the LPP.

An approximate formula may be used for Chebyshev filters

$$LA_0 = 8.686\,[N - 1.5]\,\frac{f_0}{\Delta_f Q_u}.$$

As an example, consider an eight-pole filter centered at 10 GHz. For a maximum midband attenuation of 1 dB, we find an approximate value for Q_u of 5,600.

We see that the narrower the percentage bandwidth of a filter,

achieve a specified passband insertion loss. Thus, in our example, if the passband bandwidth was 10 MHz, we would require a Q_u of 56,000. Alternatively, maintaining a Q_u of 5,600 would yield an unacceptably large midband loss of 10 dB.

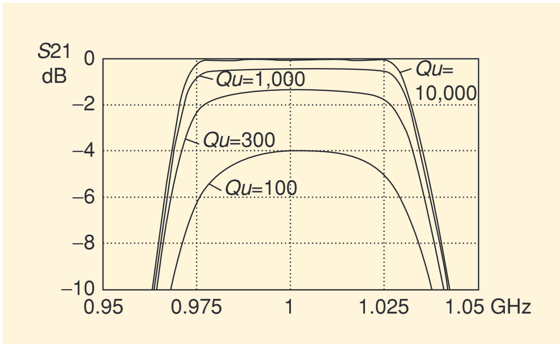

Figure 7. *Insertion loss as a function of different Q values.*

Physically, this relationship may be explained by the fact that the group delay of the filter is inversely proportional to its fractional bandwidth. Thus, for a given dissipation factor, the longer the signal stays in the network, the higher the loss. In addition, it should be noted that the group delay of a filter always increases near the passband edges. Thus, the insertion loss will normally be 1.5–2 times higher than the midband value of the passband edges. Figure 7 shows the insertion loss

ion of different Q values.

So far, we have seen that to achieve a Chebyshev bandpass filter design with a specified selectivity requirement, we need a certain number of resonant

The Chebyshev transfer function is widely used because it has a relatively high selectivity.

circuits, or resonators. The unloaded Q of these resonators determines the passband insertion loss.

In Figure 4, we see that the bandpass filter requires both series and parallel resonators, which in reality is often not convenient. This problem is solved for narrowband designs by using the inverter-coupled LPP shown in Figure 8.

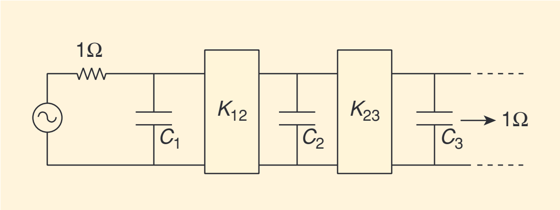

Figure 8. *Inverter-coupled LPP.*

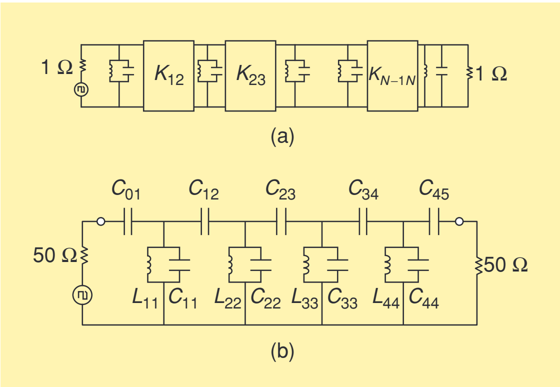

Figure 9. *(a)–(b) Coupled resonator filter.*

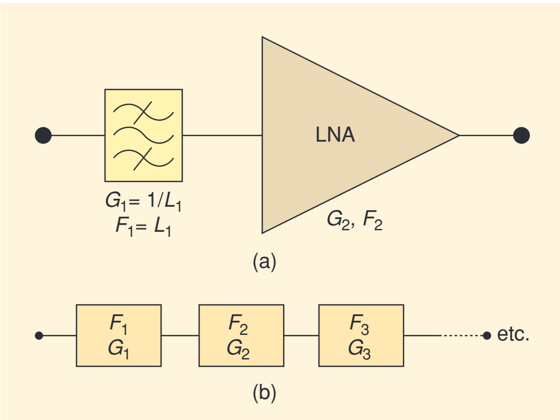

Figure 10. *(a) Receiver front-end. (b) Cascade of noisy two-ports.*

Here the inverters, denoted K_{12} etc., are 90° phase-shifting networks which eliminate the need for series inductors.

After applying the bandpass transformation to this network and approximating the inverters via capacitive pi-sections, we obtain the "coupled-resonator" bandpass filter shown in Figure 9.

This type of filter consists of a set of shunt (or series) resonators that are reactively coupled together. In reality, the Q-factor requirements on the resonators determine their physical realization. For example, lumped-element and microstrip resonators have unloaded Q factors in the low hundreds and coaxial transverse electromagnetic (TEM) resonators have Q factors in the range of 1–5,000. Waveguide and dielectric resonator realizations enable Q factors up to 50,000. Unfortunately, as a general rule, the higher the unloaded Q of a resonator, the greater its physical size. This is because, for electromagnetic resonators at least, a high Q requires lower conductor resistance and hence high surface area. It should be noted that similar arguments also apply to bandstop filters, although these are used less often.

System Considerations

In radio systems, the passband frequency response of filters affects the desired signal, and so it is essential to consider the signal attenuation relative to noise and signal amplitude and phase distortion relative to linear. In addition, the attenuation characteristic of filters is required for the two essential functions of noise bandwidth reduction and suppression of unwanted signals. Noise bandwidth has to be restricted in transmitters to avoid radiation of such signals that would interfere with other systems. In receivers, the noise bandwidth must also be reduced to the minimum level possible to maximize the received signal-to-noise ratio.

Unwanted signals occur for a variety of reasons. For example, in transmitters, where there is a well-defined baseband signal of a certain width, the task is to convert that bandwidth to some higher RF carrier frequency for radiation. The up-conversion process is necessarily nonlinear and generates unwanted signals which have to be removed. Similarly, all amplifiers are nonlinear at some elevated signal level, so filters are used where necessary to remove harmonic distortion.

In receivers, there are two quite different sources of unwanted signals. First, since reception is essentially the reverse of the transmission, the down-conversion process also produces unwanted signals that have to be suppressed. Second, there are numerous signals at the antenna, only one of which is the desired. So there is the additional complication of ensuring that antenna signals that can subsequently corrupt the desired signal are sufficiently suppressed.

It is useful to illustrate the various applications of filters and design trade-offs in different parts of a receiver system.

Bandpass filters are intended to minimize the signal attenuation in one band of frequencies, while achieving a specified attenuation in another band.

Front-End Selectivity

At the front end of a radio system, filters and amplifiers are employed to manage the many signals at different frequencies and widely varying amplitudes that are picked up by the antenna. In almost all systems, the front-end filter precedes any amplification, resulting in two basic and somewhat contradictory challenges for the filter designer:

- Minimize the loss to the desired signal
- provide sufficient attenuation to all undesired signals.

In this case, loss is important because all circuits generate thermal noise, and any attenuation prior to amplification will reduce the signal level against the fixed thermal noise background. This can be seen with reference to Figure 10(a) and the standard Fries equation for the cascade of noisy two-ports

$$F = F_1 \quad \frac{F_2 - 1}{G_1} \quad \frac{F_3 - 1}{G_1 G_2} \quad etc.$$

Considering the specific case shown in Figure 10(b), where the first two-port is a matched device with loss L_1, the gain will be $1/L_1$ and the noise figure F_1. Assuming the second stage has a noise figure F_2, substituting these values shows that the total system noise figure is $L_1 F_2$. So the loss of the first filter will directly affect the system noise figure and, hence, signal sensitivity

$$F = L_1 \qquad F_2 - 1) \, L_1 = L_1 F_2.$$

Attenuation of interference signals is also important to the clear reception of the desired signal. There are many sources of interference, but they can be subdivided into two distinct groups. First, there are strong signals (at the antenna) much larger than the desired; these can overload the sensitive input amplifiers, causing gain compression. Second, there are signals at particular frequencies, either very near to the desired or in frequency ranges of particular susceptibility. The latter devices are known as single signal spurs. Note that the problem of very strong nearby signals is very difficult to overcome, but fortunately rare because of strict global regulation of frequency bands.

Strong signals can come from high-power signals such as radar, TV, and radio broadcast transmissions, or more modest power, nearby signals such as wireless links in commercial equipment or even signals from the transmitter of the system itself, such as the case for wideband code division multiple access (W-CDMA) mobile phone signals. Even if these signals are later removed by other filters, in for example the intermediate frequency (IF) section, the effect on the desired signal is to reduce the front-end gain and therefore decrease sensitivity. Figure 11 shows a typical compression response of an amplifier. If a strong interference signal is enough to drive the amplifier into even modest compression, then that will reduce the gain for the small desired signal and hence reduce system sensitivity. This effect of a large signal is called blocking or desensitization, and as long as strong signals can be kept to no more than 10 dB below the 1-dB compression level, this can be minimized. (In practice, sometimes design compromise may only allow a few dB.) However, the levels and frequencies of such strong interferers are essentially unknown to the system designer, and it is reduction of these signals that drives the high rejection levels in very wide stopband widths above and below the desired signal frequencies. Some

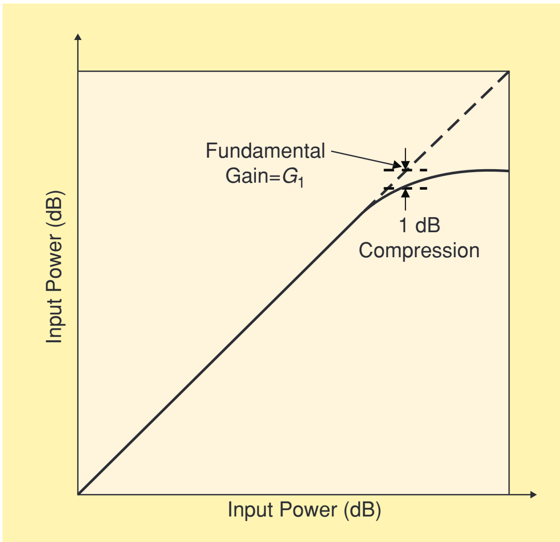

Figure 11. *Compression characteristics of amplifiers.*

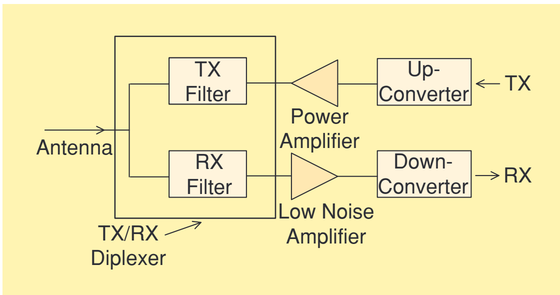

Figure 12. *Block diagram of a typical radio transceiver front end.*

Single signal spurs are a side effect of frequency mixing which is used to convert the desired RF signal to baseband for demodulation.

of the front-end filter considerations can be illustrated by the common full-duplex radio transceiver front-end shown in Figure 12.

A TX/Rx diplexer is used to duplex the transmitter and receiver onto a common antenna. The diplexer consists of a pair of transmit and receive bandpass filters connected together at the antenna port. The purpose of the receive filter is to pass the receive band with low attenuation and to attenuate unwanted interfering signals which otherwise will tend to saturate the low-noise amplifier (LNA) in the receive path. On the other hand, the transmit filter must pass the transmit band to the antenna with low attenuation and attenuate other bands to avoid the radiation of unwanted signals, including close-in spurious and harmonic terms.

As discussed earlier, the passband insertion loss of the receive filter must be sufficiently low so that the sensitivity of the receiver is not affected. On the other hand, the transmit filter is connected at the output of the power amplifier. The effect of passband attenuation in the transmit filter is to reduce the transmit power.

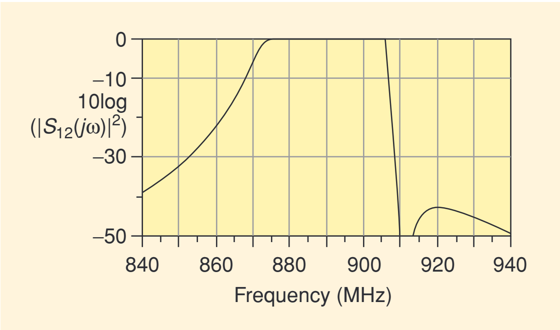

Figure 13. *Asymmetric bandpass response.*

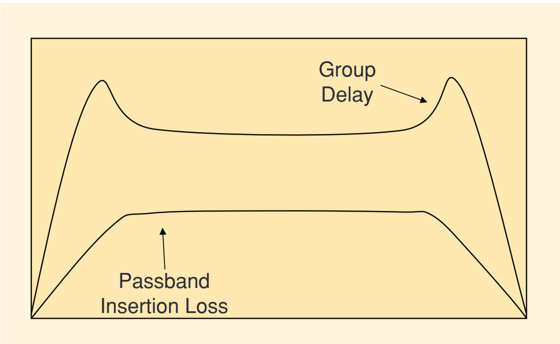

Figure 14. *Passband insertion loss and group delay.*

Naively, one could compensate for this lost power by simply increasing the power output from the power amplifier. However, in many systems, especially cellular radio, the transmit power amplifier is a complex device that has been designed to maximize both dc-RF efficiency and linearity. Attenuating its output power in the transmit filter corresponds to reducing its efficiency, and increasing its power output will reduce its linearity, producing more unwanted out-of-band distortion. Thus, excess transmit filter losses are frowned upon by systems designers. Thus, minimizing the passband insertion loss in both receive and transmit filters is of paramount importance. Typical specifications for filters in cellular radio base stations are in the range of 0.23 –l dB.

The proximity in frequency of the TX and Rx also requires considerable attenuation of the TX signal in the Rx. A mobile phone TX is typically $>$ 20 dBm, while the high sensitivity of the Rx is achieved with an LNA with an input 1-dB compression point of typically -30 dBm. This kind of requirement drives the need for relatively complicated asymmetric filter responses such as that shown in Figure 13.

Single signal spurs are a side effect of frequency mixing which is used to convert the desired RF signal to baseband for demodulation. In a mixer, for a given IF output frequency, there are two possible primary input frequencies as well as a number of other signals with lesser sensitivity. Essentially, the IF output of a mixer can be defined by the equation

$$F_{IF} = |\pm m \cdot F_{RF} \pm n \cdot F_{LO}|$$

where n and m are integers, of any value, that represent harmonics of the signals generated by the nonlinear process of mixing. For example, in a converter where the desired signal is IF = RF-LO, using the equation above, that corresponds to $m = 1$ and $n = -1$. But it is also clear that there is another possible signal called the image, at LO IF (i.e., $m = 1$ and $n = 1$) that will produce the same IF output and hence will be down-converted and demodulated in the same way as the desired signal. Image signal suppression is typically 60 dB or more and this is often seen as a distinct frequency band in the required filter attenuation characteristics of the front-end filter.

With all the different m and n integer combinations, there are clearly numerous other possible input frequencies that can produce the IF output frequency. Fortunately, the conversion loss of the mixer increases as m and n increase, so these possibilities are less problematic than the image. Because of the additional losses, there are always a finite number of combinations that have to be considered. To the system engineer, analysis of the susceptibility of different frequencies to all the mixer intermodulation possibilities is called frequency planning. Unfortunately, while the frequencies are relatively easy to predict, the mixer conversion levels for different m and n values are very dependant on the type and actual port terminations of the mixer used.

The other key issue to realize is that the conversion loss of the mixer for any m and n combination depends on the signal levels at the mixer. So for any RF signal the converted level in the IF in dB is $P_{IF} = \ K_{nm} \quad mP_{RF})$, where K_{nm} is a constant for a particular m and n combination. This means that attenuation of L dB of a single signal spur in the RF filter will result in an mL reduction of that signal in the IF. Then, since the signal follows the same mixing process, in terms of signal to interference, levels the spurious will be suppressed by $(m - 1)\,L$.

IF Response

In the IF section of a system, the desired signal is at a fixed or possibly a relatively narrow range of frequencies. Here a highly selective fixed filter is used to isolate the desired from all the nearby signals received by the antenna. Note that, by definition, single signal spurs are those that get converted to the IF frequency band and so cannot be filtered out at this stage, which is why it is vital to reduce them in the RF section. However, the IF filter will remove all the other mixing products and any local oscillator (LO) leakage. Note that the LO signal is usually quite large (typically $>\ \ 6$ dBm) and so leakage of that signal into the IF represents a significant interferer; suppression of the LO to avoid blocking of the IF amplifiers is one consideration in IF filter design. The other is the very large passband to stopbands ratio necessary to pass the wanted signal (the difference) while rejecting the unwanted signal (the sum).

On the other hand, since the IF filter passband is often only as wide as the desired modulation signal, it is important to consider passband amplitude and phase distortion through the filter. Figure 14 shows a typical passband response for an IF filter. It can be seen that towards the band edges, both the amplitude and group delay response are not constant, which produces distortion in the received signal. As most modern communications systems employ more and more complex modulation to achieve high data rates per unit bandwidth, these passband requirements become more critical.

Finally, note that in contrast to the front-end case, while the IF filter itself will have a noise figure equal to the loss, it is not usually critical because there is net gain at this point in the system. This can be understood by reconsidering Figure 10(a) and the associated noise figure (1). Assume all the stages prior to the IF filter are considered together as a net gain $G1$ and noise figure $F1$. Then if the IF filter noise figure is $F2$, the net gain $G1$ will reduce the contribution of $F2$ to the total. Hence, in the IF, while attenuation slope will distort the signal, modest constant loss can be tolerated as it hardly affects signal-to-noise ratio.

In radio systems, the passband frequency response of filters affects the desired signal.

Demodulation Bandwidth

With advancements in digital signal processing, the historic boundary between IF and baseband processing is now not well defined. Nevertheless, whether realized in the digital or analog domain, the key functionality remains the same.

Harmonic distortion from the strong signals generated from the amplification necessary for the desired signal needs to be reduced prior to demodulation. This is because demodulation is essentially a mixing process, so just like the single signal spurious issue at the front-end, such signals would otherwise mix back into the signal band.

Finally, since noise power is proportional to bandwidth, whereas the signal is not, baseband filtering is used to restrict the noise power and hence maximize signal-to-noise ratio. However, just like for an IF filter, this must be done with due care to minimize amplitude and phase distortion to the signal itself.

Conclusions

We have examined the design of microwave filters from a systems perspective. First, an overview of filter design techniques has shown the important relationships between filter selectivity, degree, bandwidth Q factor, and insertion loss. We have then examined systems issues, with particular emphasis on receivers, and considered the issues related to filtering at the front-end, IF, and baseband stages. By understanding the key principles from the component and system perspectives, designers can appreciate the trade-offs and attain better solutions to particular problems.

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