Receiver Design

Prof. Tzong-Lin Wu
EMC Laboratory
Department of Electrical Engineering
National Taiwan University

- The receiver mush be very sensitive to -110dBm and working on strong adjacent channel signals.
- Minimum detectable signal, dynamic range and the need for AGC.
- Filtering at different stage of receiver for image and spurious rejection.

Global star LEO low earth orbit satellite phone

10.1 Receiver Architecture

Receiver requirements

- High gain ~100dB
- Selectivity
- Down-conversion
- Detection of received signal
- Isolation from the transmitter to avoid saturation of receiver.

Because the typical signal power level from the receive antenna may be as low as -100 to - 120 dBm, the receiver may be required to provide gain as high as 100 to 120 dB.

This much gain should be spread over the RF, IF, and baseband stages to avoid instabilities and possible oscillation;

It is generally good practice to avoid more than about 50-60 dB of gain at any one frequency band.

Selectivity can be obtained by using a narrow bandpass filter at the RF stage of the receiver, but the bandwidth and cutoff requirements for such a filter are usually impractical to realize at RF frequencies.

It is more effective to achieve selectivity by downconverting a relatively wide RF bandwidth around the desired signal, and using a sharpcutoff bandpass filter at the IF stage to select only the desired frequency band.

Full-duplex communications systems usually use separate frequency bands for transmit and receive, thus avoiding the difficult (but not impossible) problem of isolating incoming and outgoing radiation at the same frequency.

In addition, it is often preferred to use a single antenna for both transmit and receive. In this case it is necessary to use a duplexing filter to provide isolation between the transmitter and receiver, while still providing a signal path with the antenna

Tuned radio frequency (TRF) receiver

A TRF receiver employs several stages of RF amplification along with tunable bandpass filters to provide high gain and selectivity.

But tuning is very difficult because of the need to tune several stages in parallel, and selectivity is poor because the passband of such filters is fairly broad.

In addition, all the gain of the TRF receiver is achieved at the RF frequency, limiting the amount of gain that can be obtained before oscillation occurs, and increasing the cost and complexity of the receiver.

TRF receivers are seldom used today.

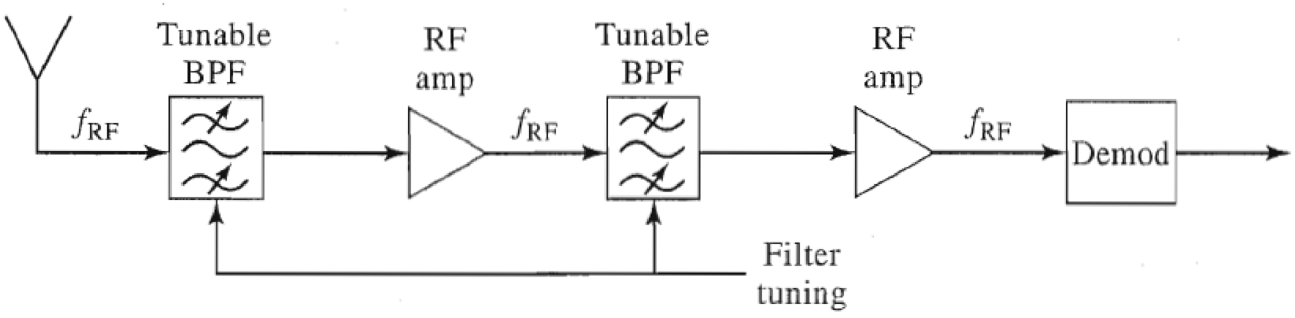


FIGURE 10.2 Block diagram of a tuned radio frequency receiver.

Direct conversion receiver (*homodyne* receiver)

Using a mixer and local oscillator to perform frequency down-conversion with a zero IF frequency.

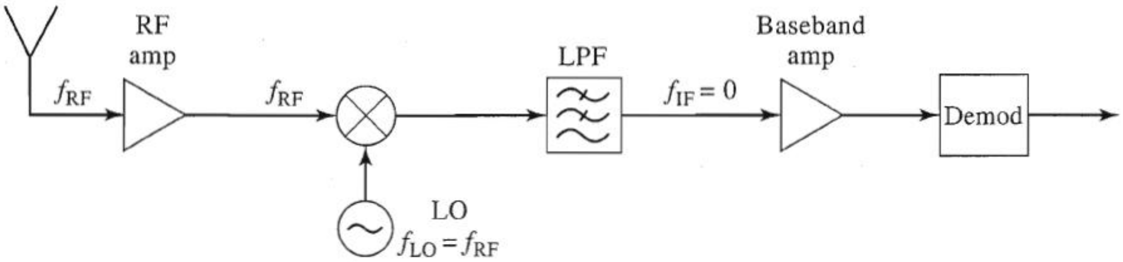


FIGURE 10.3 Block diagram of a homodyne receiver.

- Two stage amplifiers
- No need for IF amp and filter
- No need for extra circuit for AM demodulation
- No image filter required
- High stable LO source required.
- often used with Doppler radars, where the exact LO can be obtained from the transmitter, but a number of newer wireless systems are being designed with direct conversion receivers

Superheterodyne receiver

- A midrange IF allows the use of sharper cutoff filters for improved selectivity, and higher IF gain through the use of an IF amplifier.
- Tuning is conveniently accomplished by varying the frequency of the local oscillator so that the IF frequency remains constant.
- The majority of broadcast radios and televisions, radar systems, cellular telephone systems, and data communications systems

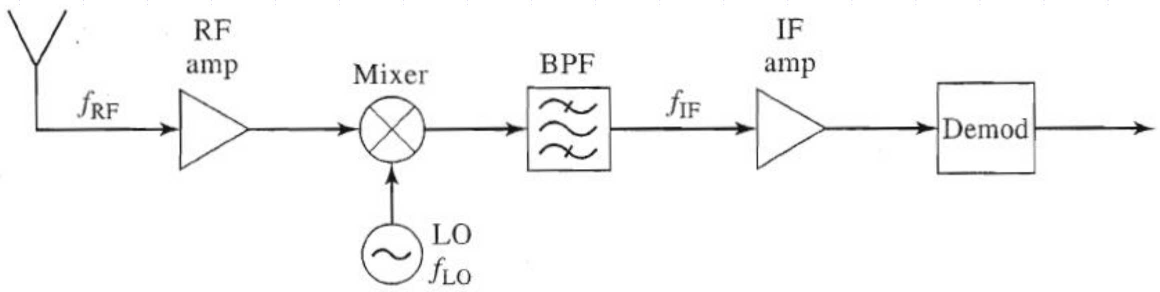


FIGURE 10.4 Block diagram of a single-conversion superheterodyne receiver.

At microwave and millimeter wave frequencies it is often necessary to use two stages of down conversion to avoid problems due to LO stability.

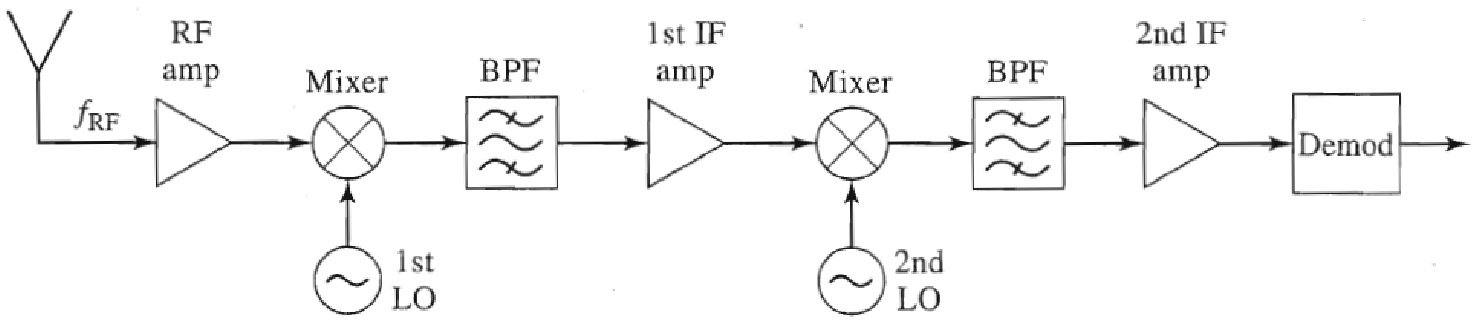


FIGURE 10.5 Block diagram of a double-conversion superheterodyne receiver.

Duplexing

If a single antenna is to be used for both transmit and receive in a duplex system, a duplexer must be used to allow both the transmitter and receiver to be connected to the antenna, while preventing the transmit signal from directly entering the receiver.

Isolation between the transmitter and receiver is usually required to be greater than 100 dB.

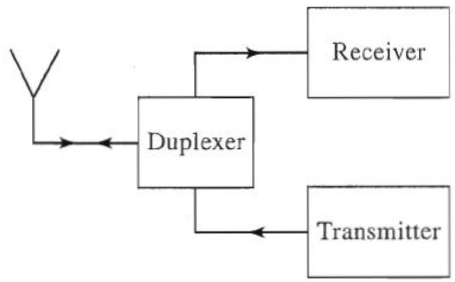


FIGURE 10.6 Use of a duplexer to connect a common antenna to a transmitter and a receiver, while providing isolation between the two.

- Tx and Rx not at the same time, a T/R switch is ok.
- Diode switch can operate at microseconds, offer 40dB isolation.
- For Tx and Rx at different frequency, bandpass filters are required for duplexer.

Full-duplex systems usually use separate transmit and receive frequency bands with bandpass filters to provide duplexing.

They can also provide some preselective filtering on receive, and attenuate spurious out-of-band signals from the transmitter. Duplexing filters often have insertion losses on the order of 1-3 dB, however, which degrades the noise figure of the receiver.

A related component is a *diplexer, a term generally used to refer to a device that* combines two or more frequency components into a single channel.

Since a duplexing filter used with different transmit and receive frequency bands fits this definition, it is sometimes referred to as a diplexing filter.

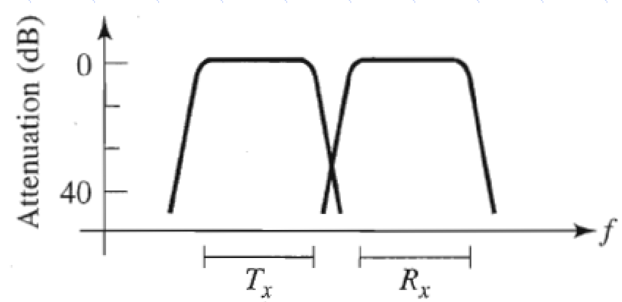


FIGURE 10.7 Transmit and receive passband responses of a duplexing filter.

EXAMPLE 10.1 FREQUENCY DUPLEXING

The AMPS and IS-54 cellular telephone system mobile units transmit over the frequency range of 824–849 MHz, and receive over the frequency range of 869–894 MHz. These two bands are each divided into 832 channels, each 30 kHz wide, to provide full-duplex communication. The first IF frequency of the receiver is 88 MHz. Compare the fractional bandwidths of the IF bandpass filter and a hypothetical tunable bandpass filter used at the RF stage. If we would like to have a minimum rejection of 50 dB between transmit and receive bands, what is the required order of the duplexing filter?

Solution

At the first IF frequency of 88 MHz, the fractional bandwidth of the 50 kHz IF filter is

$$\frac{\Delta f}{f} = \frac{0.05}{88} = 0.06\%,$$

whereas the same passband at the (midband) RF receive frequency of 882 MHz would be

$$\frac{\Delta f}{f} = \frac{0.05}{882} = 0.006\%.$$

Fractional bandwidths of 0.06% can be achieved with crystal or surface acoustic wave (SAW) filters, whereas bandwidths of 0.006% are too narrow to be achievable in practice.

To find the required order of the duplexer filter, we must first transform the transmit bandpass filter response to a low-pass prototype response. The worst-case isolation will occur for the receive frequency that is closest to the transmit band: $f = 869$ MHz. We thus require a minimum of 50 dB attenuation at 869 MHz from the bandpass filter used for the transmit band. The transmit band bandpass filter has a center frequency of

$$f_0 = \frac{824 + 849}{2} = 836.5 \text{ MHz},$$

and a fractional bandwidth of

$$\Delta = \frac{\Delta f}{f_0} = \frac{849 - 824}{836.5} = 0.03.$$

Using results from Chapter 5, the receive frequency of 869 MHz maps to a low-pass prototype (normalized to a low-pass filter with a cutoff frequency of 1 Hz) of

$$f' = \frac{1}{\Delta}\left(\frac{f}{f_0} - \frac{f_0}{f}\right) = \frac{1}{0.03}\left(\frac{869}{836.5} - \frac{836.5}{869}\right) = 2.54$$

The filter design graphs in Chapter 5 show that a filter of order $N = 4$ is required for 50 dB attenuation. ○

10.2 Dynamic Range

Minimum detectable signal (MDS)

Reliable communication requires a receive signal power at or above a certain minimum level, which we call the minimum detectable signal (MDS).

Minimum detectable signal (MDS) determines the minimum SNR at the demodulator for a given system noise power.

TABLE 10.1 Typical Minimum SNR for Various Applications

System	SNR (dB)
Analog voice	5–10
Analog telephone	25–30
Analog television	45–55
AMPS cellular	18
AM-PCM	30–40
QPSK ($P_e = 10^{-5}$)	10

SINAD: signal-plus-noise-plus-distortion to noise-plus-distortion ratio.

The ratio of (a) total received power, i.e., the received signal-plus-noise-plus-distortion power to (b) the received noise-plus-distortion power.

$$\mathrm{SINAD} = \frac{S+N}{N} = 1 + \frac{S}{N}. \tag{10.1}$$

Knowing the minimum SNR or SINAD and the noise characteristics of the receiving system allows us to calculate the minimum detectable signal power.

Minimum Detectable Power

the output signal power

$$S_o = G S_i, \tag{10.2}$$

Receiver Noise temperature

$$T_e = (F - 1)T_0.$$

total output noise power

$$N_o = kBG(T_A + T_e), \tag{10.3}$$

where B is the bandwidth of the receiver (usually set by the IF bandpass filter).

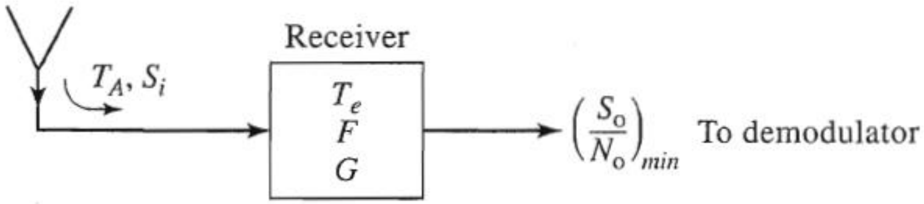


FIGURE 10.8 Block diagram of receiving system for the determination of minimum detectable signal.

Minimum detectable input signal level

$$S_{i_{\min}} = \frac{S_{o_{\min}}}{G} = \left(\frac{N_o}{G}\right)\left(\frac{S_o}{N_o}\right)_{\min} = kB(T_A + T_e)\left(\frac{S_o}{N_o}\right)_{\min}$$
$$= kB[T_A + (F-1)T_o]\left(\frac{S_o}{N_o}\right)_{\min} \quad (10.4)$$

This is an important result, relating the minimum detectable signal power at the input of the receiver to the noise characteristics of the receiving system and the minimum SNR required for that application.

This equation provides the interface between the radio link equation (e.g., the Friis equation or ground reflection link equation) and the SNR or error rate equations of Chapter 9, thereby allowing characterization of the complete wireless system.

For digital modulation, recall from (9.105) that the bit energy to noise power spectral density, E_b/n_0, is related to the SNR and bit rate R_b as follows:

$$\frac{E_b}{n_0} = \frac{S_o}{N_o}\frac{B}{R_b}. \qquad (10.5)$$

For the special case where $T_A = T_0$, (10.4) reduces to the following result:

$$S_{i_{\min}} = kBT_0F\left(\frac{S_o}{N_o}\right)_{\min}. \qquad (10.6)$$

This equation can be conveniently expressed in dB:

$$\boxed{\begin{aligned} S_{i_{\min}}(dB) &= 10\log(kT_0) + 10\log B + F(dB) + \left(\frac{S_o}{N_o}\right)_{\min}(dB) \\ &= -174\,\text{dBm} + 10\log B + F(dB) + \left(\frac{S_o}{N_o}\right)_{\min}(dB) \end{aligned}} \qquad (10.7)$$

Two important points:

Although (10.6 and10.7) are sometimes used in place of (10.4), it is important to realize that (10.6 and 0.7) are only valid when the antenna temperature is 290 K, this situation is seldom true in practice.

In either case, note that the minimum detectable signal level does not depend on the gain of the receiver, since both signal and noise are increased equally.

EXAMPLE 10.2 MINIMUM DETECTABLE SIGNAL

An IS-54 PCS telephone system uses QPSK with a bit rate of 46.6 kbps. The receiver has a bandwidth of 30 kHz, and a noise figure of 8 dB. The receive antenna has a noise temperature of 900 K. If the bit error rate is 10^{-5}, find the minimum detectable signal level.

Solution

Assuming Gray coded QPSK, $E_b/n_0 = 10\text{ dB} = 10$. Then the minimum SNR is found from (10.5) as

$$\left(\frac{S_o}{N_o}\right)_{\min} = \frac{R_b}{B}\frac{E_b}{n_0} = \frac{46.6 \times 10^3}{30 \times 10^3}(10) = 15.5$$

Using (10.4) gives the MDS as

$$\begin{aligned} S_{i_{\min}} &= kB[T_A + (F-1)T_o]\left(\frac{S_o}{N_o}\right)_{\min} \\ &= (1.38 \times 10^{-23})(30 \times 10^3)[900 + (6.3-1)(290)](15.5) \\ &= 1.57 \times 10^{-14}\text{ W} = 1.57 \times 10^{-11}\text{ mW} = -108\text{ dBm}. \end{aligned}$$

This value indicates the need for a total receiver gain on the order of 100 dB.

As a comparison, the (incorrect) use of (10.7) gives

$$
\begin{aligned}
S_{i_{\min}} &= -174 \text{ dBm} + 10 \log B + F(dB) + \left(\frac{S_o}{N_o}\right)_{\min} (dB) \\
&= -174 + 10 \log(30 \times 10^3) + 8 + 15.5 \\
&= -105.7 \text{ dBm}.
\end{aligned}
$$

Note that this result is in error by several dB.

Sensitivity

Receiver voltage sensitivity or sensitivity:

The minimum detectable signal power can be converted to a minimum detectable signal voltage, for a given receiver input impedance.

$$V_{i_{\min}} = \sqrt{2Z_0 S_{i_{\min}}} \cdot \ \mathrm{V\,(rms)}. \qquad (10.8)$$

Dynamic range

Receiver dynamic range

$$DR_r = \frac{\text{maximum allowable signal power}}{\text{minimum detectable signal power}}. \qquad (10.9)$$

Depends on noise, modulation scheme, and required minimum SNR.

The maximum allowable signal power could alternatively be defined by the third-order intercept point, P_3, at the input to the receiver, as this would be the maximum input power before intermodulation distortion becomes unacceptable.

EXAMPLE 10.3 DYNAMIC RANGE

Consider the AMPS cellular system, transmitting from the base station at 880 MHz with a transmit power of 20 dBm. If the transmit and receive antennas have gains of 1 dB, find the receive power versus distance under the assumption of free-space conditions. If we assume the minimum detectable signal level from Example 10.2 of –108 dBm, and a minimum distance of 10 m between the transmit and receive antennas, what is the required dynamic range of the receiver? If the third-order intercept point at the input of the receiver is –15 dBm, how close can the receiver be to the transmit antenna before third-order intermodulation distortion becomes severe?

Solution

The Friis equation can be used to calculate the received power level versus distance R between the transmitter and receiver:

$$P_r = \frac{G_r G_t P_t \lambda^2}{(4\pi R)^2}.$$

The antenna gains are $G_r = G_t = 1.26$, and the wavelength is $\lambda = 300/880 = 0.341$ m. The received power versus range is listed in the next table:

R (m)	P_r (dBm)
0.5	−3.3
1	−9.3
10	−29.3
100	−49.3
1000	−69.3
10,000	−89.3

Observe that the received signal level decreases by 6 dB with a doubling in range, and decreases by 20 dB for an increase of 10 in the range. (In the case of ground reflections, the received signal level would decrease by 12 dB for a doubling in range.)

If we limit the minimum distance between transmitter and receiver to 10 m, we see that we can expect a maximum received power level of about −29.3 dBm. For a minimum detectable signal level of −108 dBm, this gives a receiver dynamic range of

$$DR_r = (-29.3 \text{ dBm}) - (-108 \text{ dBm}) = 78.7 \text{ dB}.$$

This table shows that when $R = 1$ m, the received power is about −9 dBm. Doubling that distance to $R = 2$ m will decrease the received power to about −15 dBm, which is the level of the third-order intercept at the input to the receiver. ○

Automatic gain control

The above examples show the need for about 80-100 dB of receiver gain to raise the minimum detectable signal to a usable level of approximately 10 mW (about 1 V peak at 50 ohm**).**

Most of the gain occurring at the IF stage because

Amplifiers and other components are generally cheaper at lower frequencies.

High input signal levels may exceed the 1 dB compression point (P_1),or the third-order intercept point (P_3), of the front-end components if the gain of the early stages is too high.

Moderate level of gain at the RF stage sets a good noise figure for the receiver system.

Automatic gain control

Dynamic range at the output of the receiver is much smaller than the 80-100 dB dynamic range at the receiver input. (why?)

At the output of the receiver the detected baseband signal often drives a digital signal processing (DSP) circuit, or a digital to analog converter PAC), where the input voltage range is typically 1 mV to 1 V.

For example, in a digital PCS telephone receiver the input signal is demodulated to recover digitized data, and then converted to an analog voice signal with a DAC.

A 10-bit DAC with a maximum output voltage of 1 V has a resolution of $1/2^{10} = 1/1024 \sim 1$mV, and provides a dynamic range of 201og 1000 = 60 dB.

Automatic gain control

- The power gain through the receiver must therefore vary as a function of the input signal strength in order to fit the input signal range into the baseband processing range, for a wide range of input signal levels.

  This variable-gain function is accomplished with an automatic gain control (AGC) circuit. AGC is most often implemented at the IF stage,

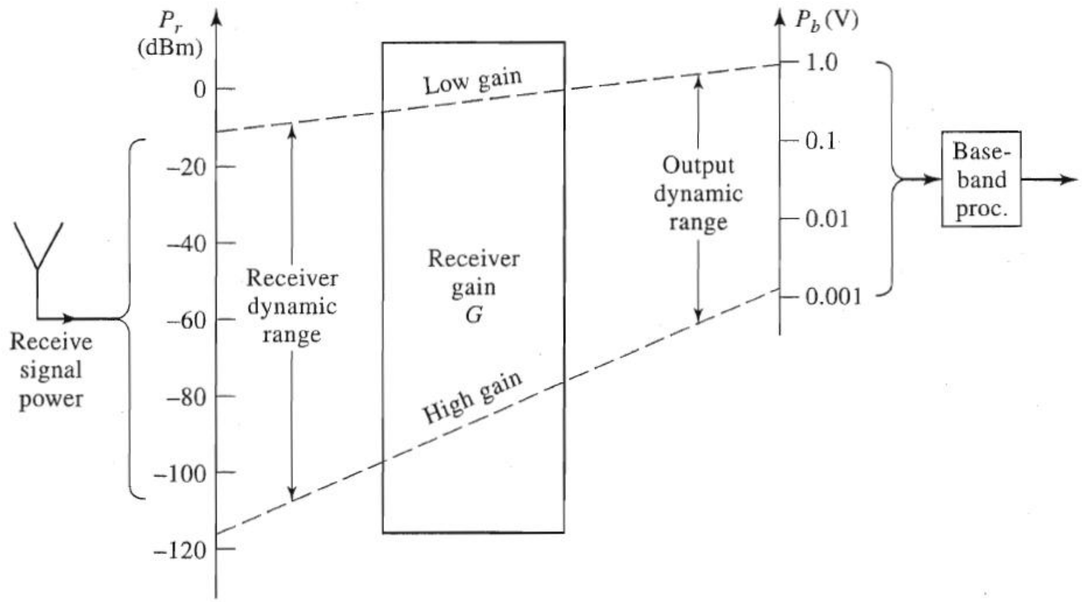


FIGURE 10.9 Diagram illustrating the change in power levels between the input and output of a typical receiver.

It consists of a variable voltage controlled attenuator (or variable gain amplifier) with a detector to convert a sample of the IF voltage to a DC value.

The rectified signal is then compared with a reference level, and passed through a low-pass filter to provide a time constant long enough to avoid having the AGC following low-frequency components of the modulated signal.

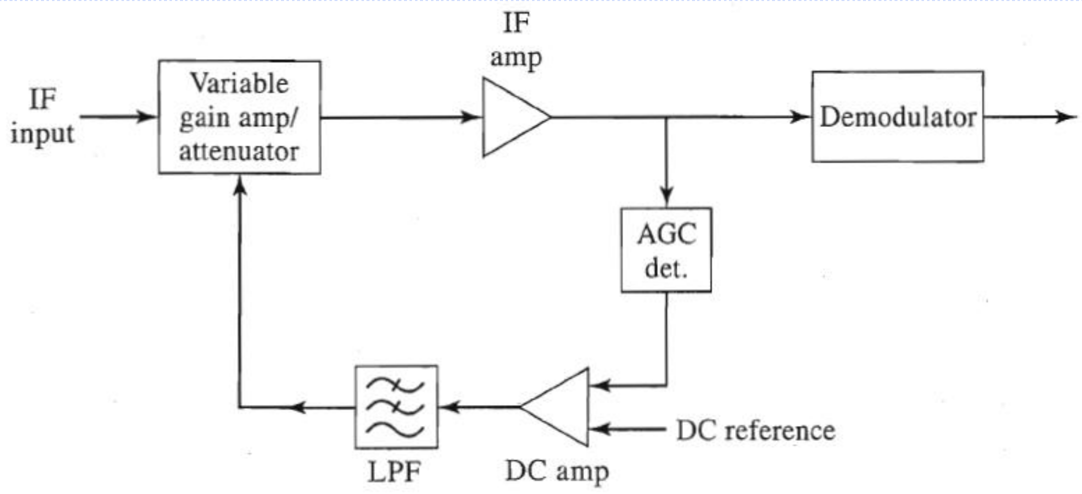


FIGURE 10.10 Block diagram of an automatic gain control circuit at the IF stage.

Compression and Third-order Intermodulation

It is important to track power levels through the stages of the receiver to ensure that P1 and P3 are not exceeded.

Power exceed P1 dB will cause harmonic distortion
Power exceed IP3 will cause intermodulation distortion

Placing an AGC attenuator in the RF stage may reduce the possibility of saturating the RF amplifier with a large input signal, but will degrade the noise performance of the receiver, even for a low attenuator setting.

$$S_{i_{\max}} + G_{\mathrm{RF}} < P_1, \tag{10.10}$$

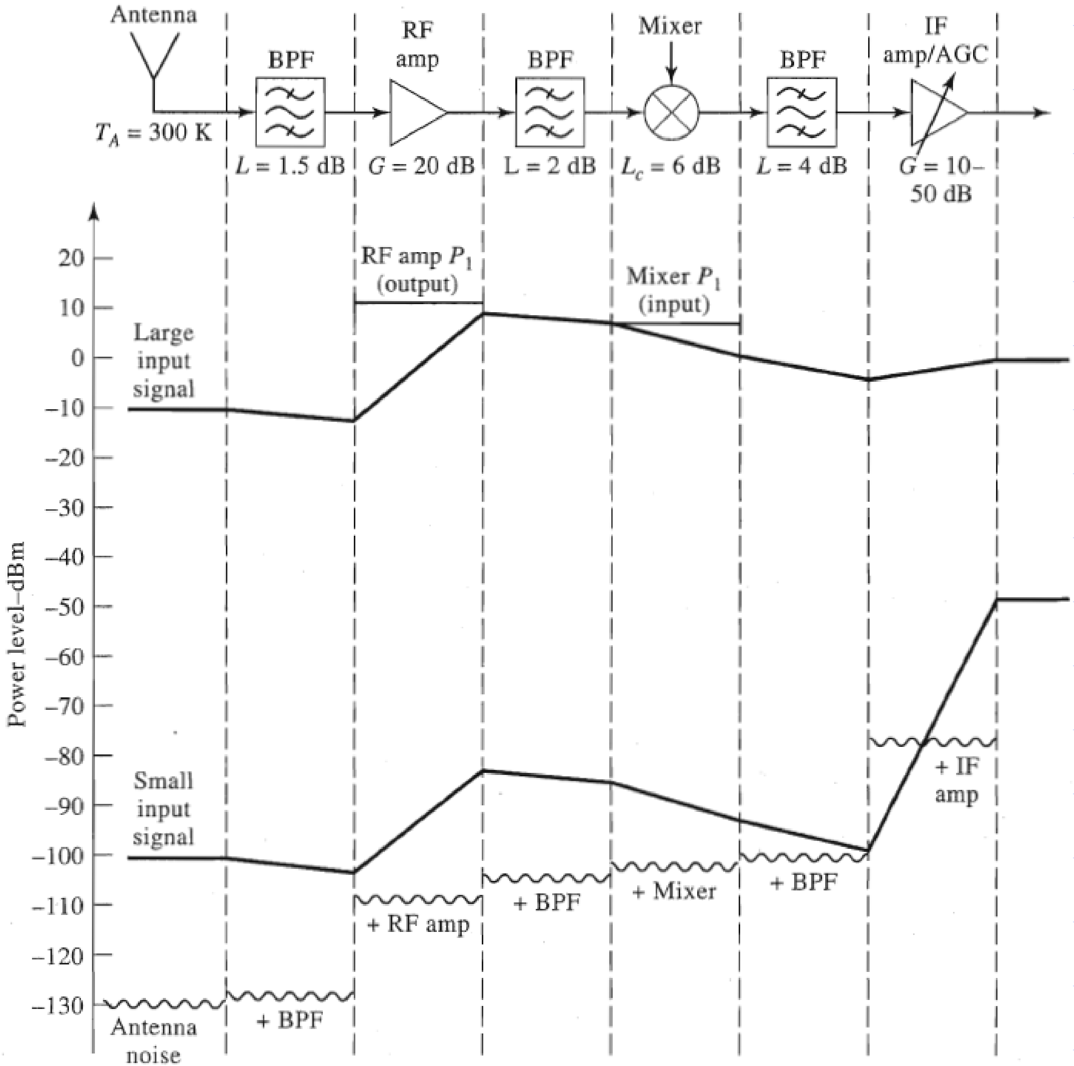


FIGURE 10.11 Diagram of power and noise levels at consecutive stages of a receiver.

10.3 Frequency conversion and filtering

A key decision in the design of a superheterodyne receiver is the choice of IF frequency. The IF frequency is related to the RF and LO frequencies by

$$f_{IF} = |f_{RF} - f_{LO}|. \quad (10.11)$$

While it is possible to use a local oscillator either above or below the RF signal frequency, most receivers use the lower sideband, so that the LO frequency is given by

$$f_{LO} = f_{RF} - f_{IF}, \quad (10.12)$$

The mixer also responds to an RF image frequency separated by twice the IF frequency:

$$f_{IM} = f_{RF} - 2f_{IF}. \quad (10.13)$$

Because the image signal is often removed by filtering, using a large IF frequency eases the cutoff requirements of the image filter. In addition, to ensure that the image frequency is outside the RF bandwidth of the receiver, it is necessary to have

$$f_{IF} > \frac{B_{RF}}{2}, \quad (10.14)$$

where B_{RF} is the RF bandwidth of the receiver.

IF frequency less than 100MHz for lower cost of components.
Possible selections are: crystal, ceramic and surface acoustic wave filters

Filtering

Filtering is required in a superheterodyne receiver to provide interference rejection, image rejection, selectivity, and suppression of LO radiation.

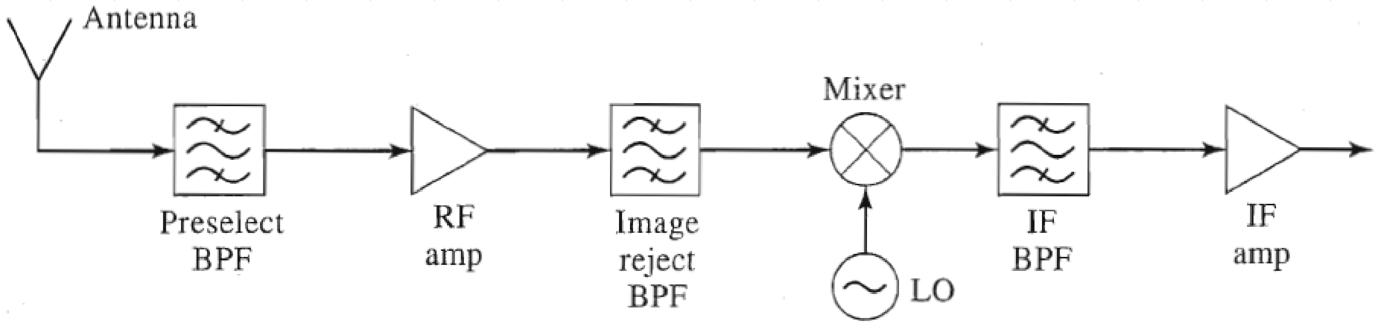


FIGURE 10.12 RF and IF filtering in a superheterodyne receiver

A preselect filter

This is a bandpass filter set to the RF tuning range of the receiver.

It rejects out-of-band interference, which is particularly important for preventing strong interference signals from saturating the RF amplifier or mixer.

In order to keep the noise figure as low as possible, this filter should have a low insertion loss. This implies that its cutoff characteristics will not be very sharp, so this filter generally does not provide much rejection of the image frequency.

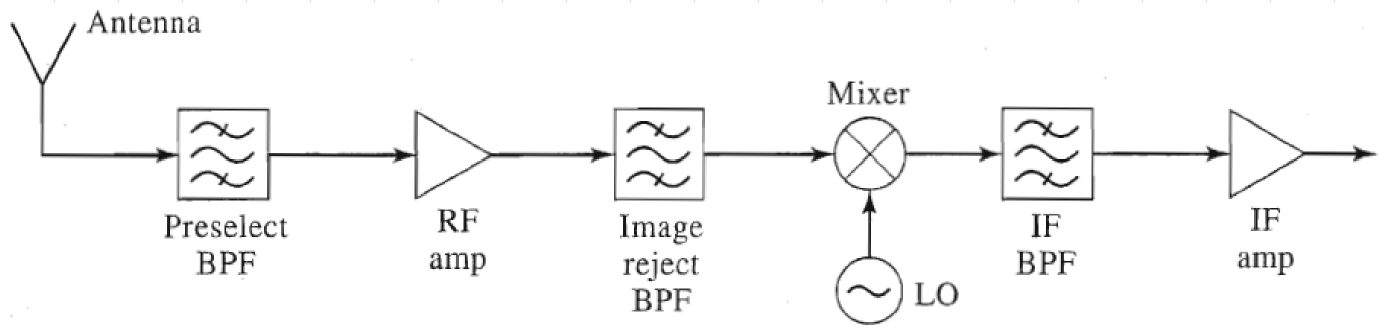


FIGURE 10.12 RF and IF filtering in a superheterodyne receiver

Usually placed after the RF amplifier, where the insertion loss associated with a filter having a sharp cutoff will have less effect on the noise figure of the receiver.

This filter is often a ceramic dielectric resonator type. The image reject filter may also reduce the effect of possible harmonic distortion from the RF amplifier.

LO Radiation (Leakage)

Because the local oscillator frequency is separated from the RF frequency only by the IF frequency, it often lies in the RF passband of the receiver, and may pass back through the RF stages to be radiated by the antenna.

Such radiation may interfere with other users, and therefore must be attenuated to a very low level. This is usually accomplished by the combined attenuation of the preselect and image reject filters, the LO-RF isolation of the mixer, and the reverse attenuation of the RF amplifier.

Because the LO is only one IF frequency away from the RF frequency, while the image is twice the IF frequency away, it is sometimes more difficult to meet the requirement for low LO radiation than it is for image rejection.

The IF bandpass filter sets the overall noise bandwidth of the receiver, as well as removing most unwanted mixer products such as

many intermodulation products of the form $nf_{LO} \pm mf_{RF}$.

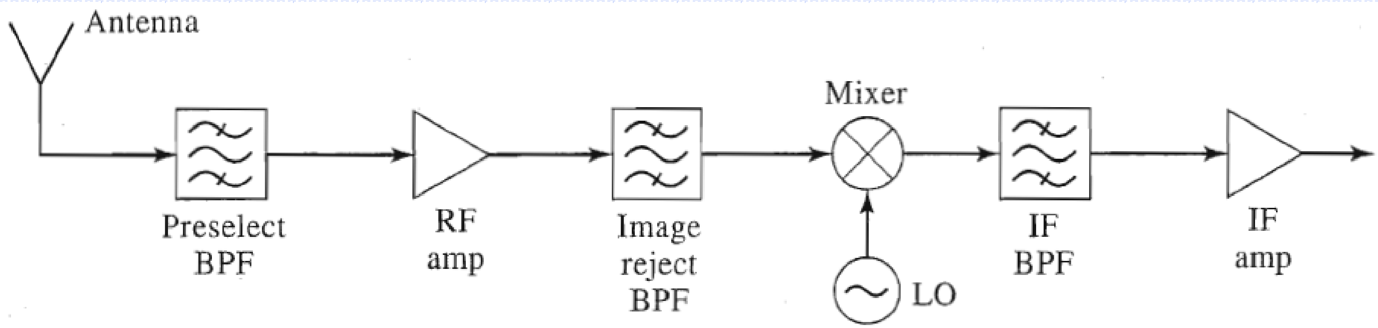


FIGURE 10.12 RF and IF filtering in a superheterodyne receiver

Spurious-free range

The nonlinear action of the mixer produces the sum and difference frequencies of the input signals, along with smaller levels of power at the intermodulation products at

$$f = |mf_{\mathrm{RF}} - nf_{\mathrm{LO}}|, \qquad (10.15)$$

Most of these products are far outside the passband of the IF stage, but some may fall within the IF band.

These are called *spurious responses* (or "spurs"), and are a problem because the receiver will respond to undesired signals at RF frequencies within its tuning range that produce spurs within the IF passband.

It is usually sufficient to specify that the order of spurious responses within the IF passband be greater than a value in the range of 6-10.

In order to check the spurious response, the following procedure can be used. Because of the large number of combinations to check, it is best to write a simple computer program to evaluate the intermodulation product frequencies.

- Divide the RF tuning range of the receiver into K frequencies, spaced by the IF bandwidth.
- For each RF frequency, compute the required LO frequency from (10.12).
- Compute the intermodulation frequency using (10.15), for $1 \le m \le M$ and $1 \le n \le N$, where M and N are the upper limits of the maximum order to be searched.
- A spurious response lies within the IF passband if the value of (10.15) is less than the IF bandwidth

The number of combinations that must be checked is $K \times M \times N$.

While (10.15) gives all possible spurious frequency products, mixers often will inherently suppress some products due to symmetries and phase cancellations in the mixer circuit.

Double-balanced mixers, for example, will reject all spurious responses where either m or n is even. Depending on their design, singly balanced mixers may reject some, but not all, products with m or n even.

EXAMPLE 10.4 SPURIOUS RESPONSES

Consider an AMPS receiver operating from 869 to 894 MHz, with a first IF of 88 MHz, and an IF bandwidth of 50 kHz. Find the possible spurious responses of order less than 10.

Solution

A short computer program was written to implement the above procedure, checking over 50,000 combinations in less than one second. The following products resulted in frequencies within the IF passband:

f_{RF} (MHz)	f_{LO} (MHz)	m	n	$\lvert mf_{RF} - nf_{LO}\rvert$ (MHz)
all	all	1	1	88
880.	792.	8	9	88

The first row, with $m = n = 1$, occurs for any RF frequency in the range from 869 to 894 MHz, and represents the desired down-conversion to the IF. The second row represents a spurious response with $m = 8$ and $n = 9$. This product may be suppressed if a doubly balanced mixer is used. ○

10.4 Example of practical receivers

FM broadcast receiver

- 88MHz~108MHz with 200kHz channel spacing.
- Each radio station has ±75kHz with sensitivity at 10μV
- IF at 10.7MHz.

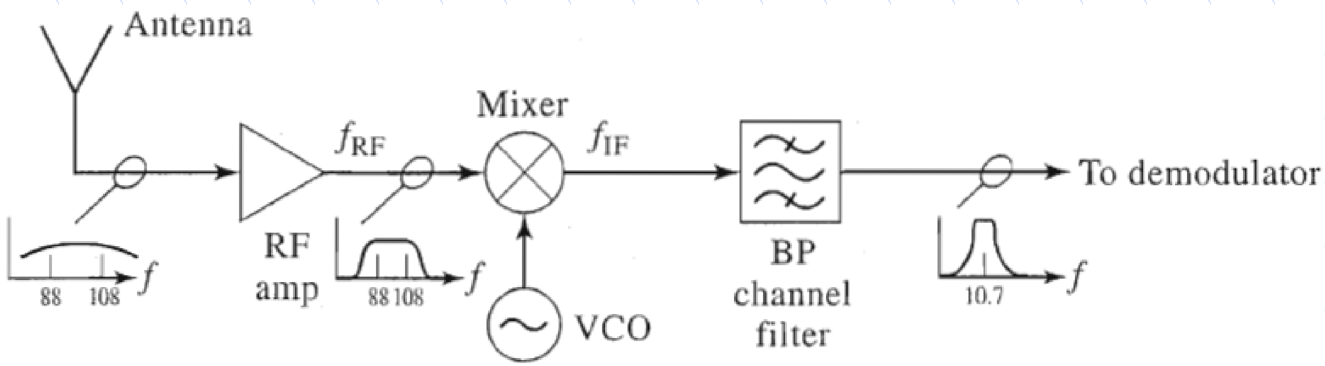


FIGURE 10.13 Block diagram of an FM broadcast receiver.

the RF amplifier usually provides a small amount of bandpass filtering to reject signals far outside the FM band.

The desired channel cannot be tuned at the RF stage, however, because the required tunable filter bandwidth of 150 kHz/100 MHz = 0.15% cannot be inexpensively realized in practice.

Selectivity is then provided by the IF bandpass filter, with a fixed center frequency of 10.7 MHz. The fractional bandwidth of this filter is about 150 kHz/l0.7 MHz = 1.4%, which can be easily realized.

- f_{LO} at 77.3~97.3MHz, tuning ratio=97.3/77.3=1.3
- Image at 66.6~86.6MHz, outside of FM band.
- The image frequency will always lie outside the receiver band if the IF frequency is selected to be at least (108 - 88)/2 = 10 MHz.
- RF, IF amplifier, VCO and mixer are integrated into a single chip.
- IF bandpass filter is hard to integrate.

Engineers at Philips in the 1980s to propose the innovative idea of lowering the IF frequency from 10.7 MHz to 70 kHz.

This converts the desired FM channel down to baseband (0 to 70 kHz), so channel selectivity can be obtained with a low-pass filter, which can be easily implemented with active integrated circuitry.

The image frequency in this case is located 2 f_{IF} = 140 kHz away from the desired RF channel, placing it between the selected channel and the next lower channel.

The image signal therefore is not attenuated by the RF amplifier or preselect filter (if one is used).

The noise from the image bandwidth will reduce the receiver SNR by 3 dB. This is a trade-off in performance for a very high degree of miniaturization-FM receivers of this type can be made as small as a wristwatch.

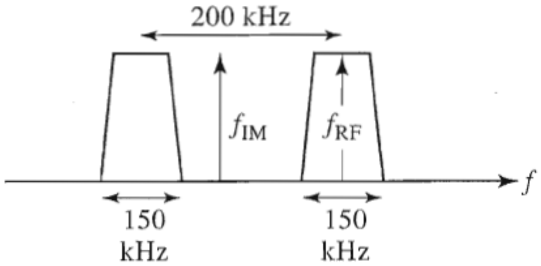


FIGURE 10.14 Spectrum of two FM channels and the image frequency for an FM receiver with a 70 kHz IF frequency.

Digital cellular receiver

- IS-54 at north America.
- Tx at 824-849 MHz , Rx at 864-894 MHz
- 30kHz channel width with 832 full duplex FDM.
- Same at AMPS system
- AMPS: analog FM
- IS-54: QPSK with 48.6kbps and 3 users share the same channel by TDMA

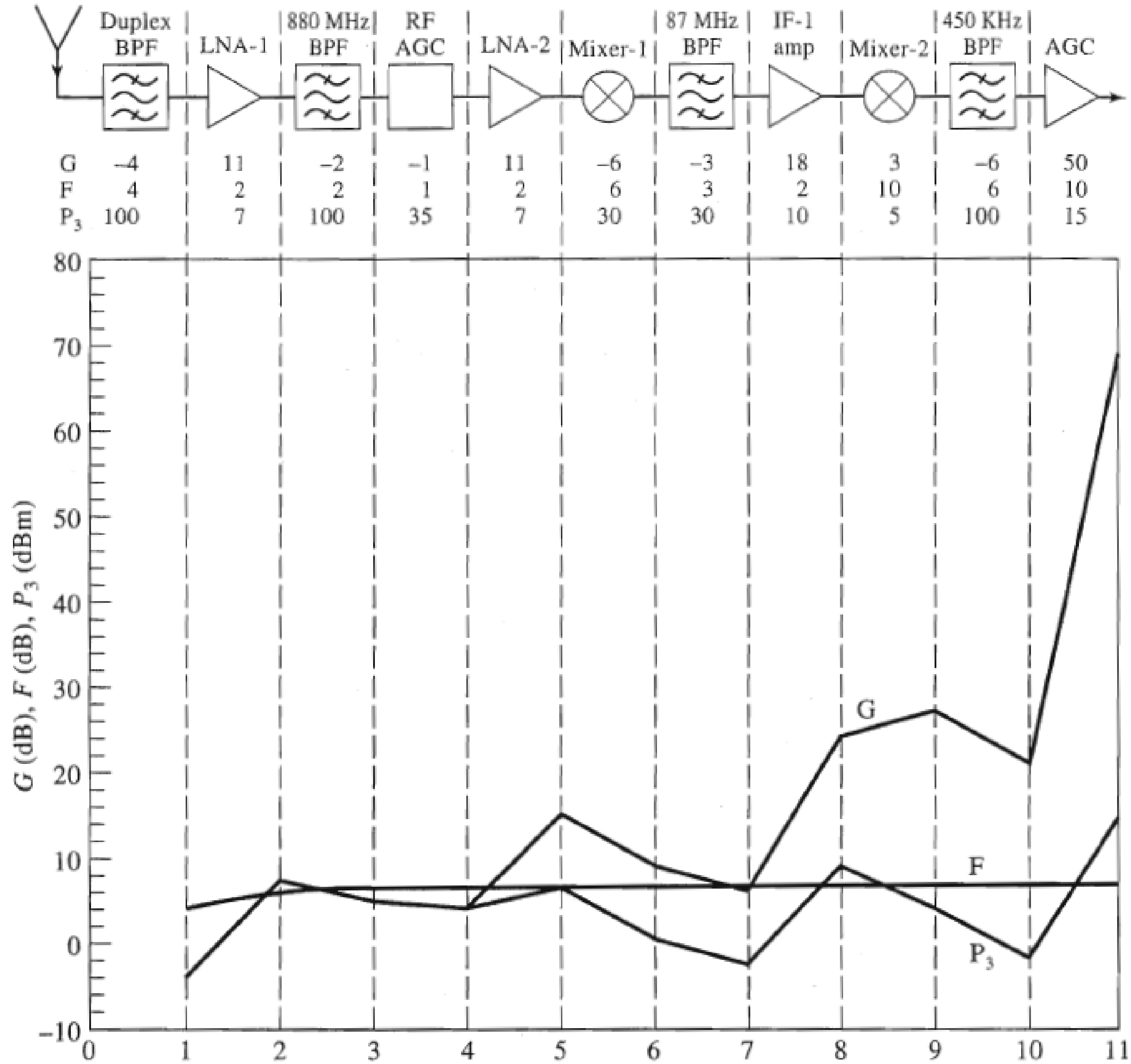


FIGURE 10.15 Block diagram of an IS-54 digital PCS receiver. The gain (in dB), noise figure (in dB), and the third-order intercept point (in dBm, reference at output) are listed for each component. The cascaded gain, noise figure, and third-order intercept are plotted progressively through the system.

Image at (869-894)-2(87)= 695-720MHZ

Millimeter Wave Point-to-Point Radio Receiver

- Base station to MTSO (mobile telephone switching offices) and PSTN
- High data rate and good reliability required.
- More economic then fiber or coaxial cable.

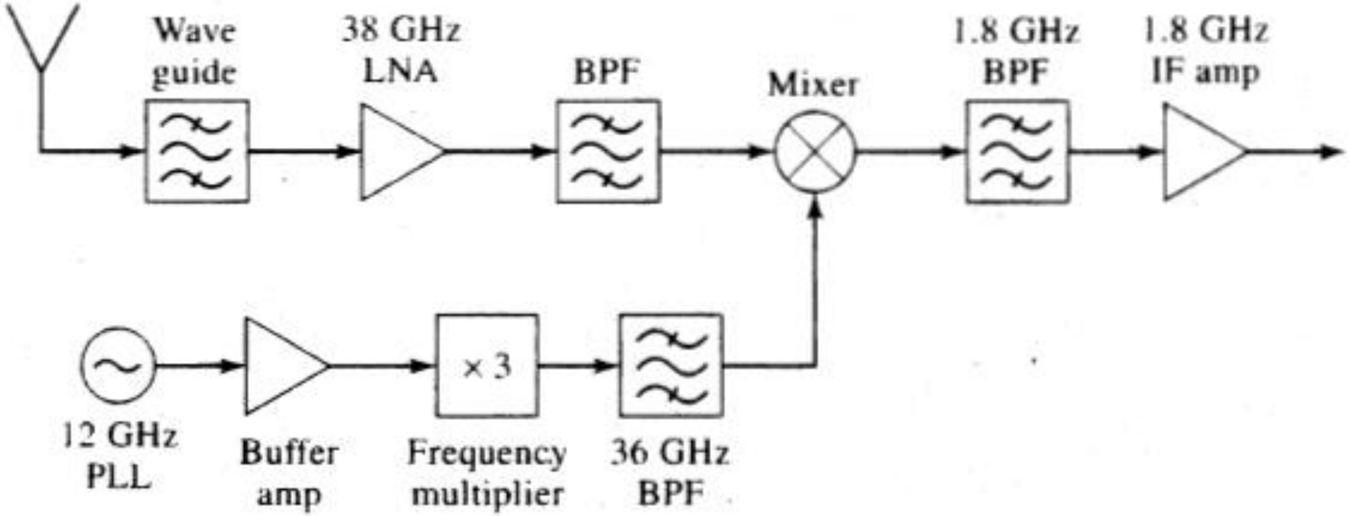
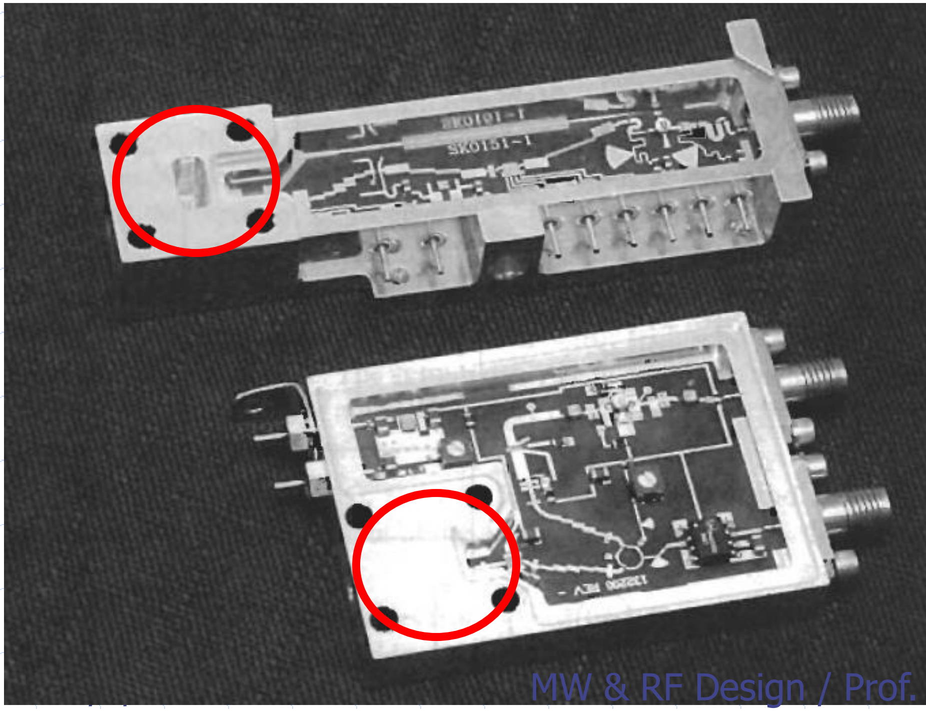


FIGURE 10.16 Block diagram of the front-end of a 38 GHz point-to-point radio receiver.



Tx

Rx

components are:

38 GHz waveguide transition:	insertion loss = 1.0 dB
38 GHz low-noise amplifier:	gain = 20 dB noise figure = 3.5 dB third-order intercept = 15 dBm
38 GHz band pass filter:	insertion loss = 4 dB
first mixer:	conversion loss = 7 dB noise figure = 7 dB third-order intercept = 10 dBm
1.8 GHz IF amplifier:	gain = 13 dB noise figure = 2.5 dB third-order intercept = 25 dBm

Application of the cascade formula gives the following overall characteristics of the receiver front end:

receiver gain:	21 dB
receiver noise figure:	4.9 dB
receiver intercept point:	−5.5 dBm (at input)

The 36 GHz local oscillator has a phase noise level that is 85 dB below the carrier at an offset of 100 kHz.

Direct conversion GSM receiver

- Direct conversion eliminated the need of IF filters, amplifiers and IF LOs.
- Tx at 880-915 MHz and Rx at 925-960 MHz.
- Tx and Rx are multiplexed by TDMA, therefore only a T/R switch is required for duplexing.

Direct conversion GSM receiver

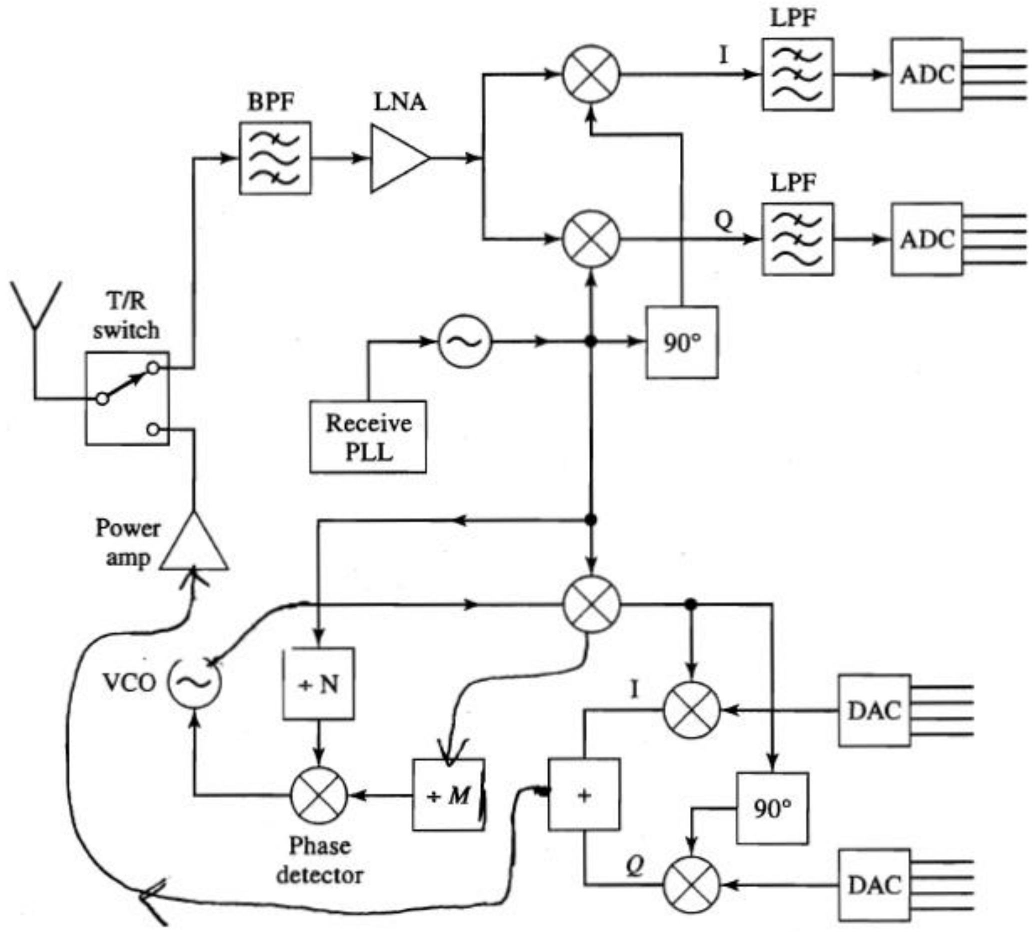


FIGURE 10.18 Block diagram of a GSM transceiver front end, consisting of a transmitter and a direct conversion receiver.

Wireless Standards

*IMS2007 TSA-3

Feature	GSM / GPRS EDGE	WCDMA	802.11(abg)	802.11n	802.16e WiMax	802.15.3a UWB	4G (TBD)
Defined by	ETSI	3GPP/ETSI	IEEE	IEEE	IEEE	IEEE	TBD
Access	FDM/TDMA	CDMA / Full Duplex	CSMA-CA	CSMA-CA	TDMA	CSMA-CA	TBD
Modulation	0.3 GMSK (1bit/Symbol)	QPSK	DBPSK DQPSK 16QAM 64QAM	64 QAM on 108 OFDM	Fixed: OFDM Mobile: OFDMA : BPSK, QPSK, 16QAM, 64QAM	OFDM/ PPM	MC-CDMA OFDM Flash-OFDM
Modulation Filter	0.3 Gaussian	Root Raised cosine a=0.22	Gaussian or vendor specific	TBD	Depends on format	Dependent on PPM or OFDM	TBD
Chan. Spacing	200(kHz)	5MHz	Depending on region and flavor. Chan. BWs include 10 MHz, 25 MHz, 30 MHz	20MHz or 40MHz (yet TBD)	Fixed: 1.25 – 20 MHz Mobile: 1.25, 5, 10MHz	North America > 500 MHz for 3.1-10.6 GHz	TBD
Symbol or Bit Rate	270.8bks/sec	3.84 Mcps	11 Mchips/sec	Above 100Mbps	Up to 87 ksps	Depends on Version	Above 100Mbps
RF Frequencies (MHz)	GSM900 880-915 DCS1800 1710-1785 PCS1900 1850-1910 (RX) 925-960 180	Band(I,II,III, & V) I : 1920-1980(TX) I : 2110-2170(RX) II : 1850-1910(TX) II : 1930-1990(RX) III : 1710-1785(TX) III : 1805-1880(RX) V : 824-849(TX) V : 869-894(RX)	802.11b 2.4-2.48 GHz 802.11a & g 4.9-5 GHz (Japan) 5.03-5.091 GHz (Japan) 5.15-5.35 GHz (UNII) 5.47-5.75 GHz 5.725-5.85 GHz (ISM, UNII)	802.11b 2.4-2.48 GHz 802.11a & g 4.9-5 GHz (Japan) 5.03-5.091 GHz (Japan) 5.15-5.35 GHz (UNII) 5.47-5.75 GHz 5.725-5.85 GHz (ISM, UNII)	Fixed < 11GHz Mobile : 2 – 6 GHz	North America > 500 MHz for 3.1-10.6 GHz	Initial Proposal Suggest 2GHz – 8GHz
Approx. Range	~5km	~5km	~100m	~300m	As high as 30km	3m	~5km TBD
Peak TX Power	33dBm (MS)	24dBm	20dBm(bg) Country specific	TBD	~1Watt: fixed (TBD) ~21dBm: mobile (TBD)	-41dBm/Hz (400	TBD
Key Hardware & Performance Challenges	(RX) Blocking @ 3MHz (RX) (TX) TX noise in RX band @ 20MHz offset from carrier	TX-to-RX leakage @ 20MHz from carrier. Creates IP2, IP3 and NF issues.	Modulation of 64 QAM places high linearity demand on TX & RX, in addition to challenging synthesizer performance.	Cross channel interference.	Large signal BW (up to 20MHz) requires low EVM, integrated phase from 10Hz-10MHz must be -36dBc	Difficult to meet switching speeds for MBOA.	TBD
EVM	5°(RMS), 20°(peak)	17.5%	5%(RMS)	TBD	3% (TBD)		TBD
Synt. Settling Time	180µsecs					MBOA 9.47ns	TBD
Websites	www.etsi.org www.3gpp.org	www.3gpp.org	www.wi-fi.org www.ieee802.org/11	www.wi-fi.org	www.wirelessman.org	www.ieee802 .org/15	TBD

Jan. 5, 2008