CornellEngineering
Electrical and Computer Engineering

ECE 4880: RF Systems

Fall 2016

Chapter 2: Overview of RF components and system

Reading Assignments:

1. T. H. Lee, *The Design of CMOS Radio Frequency Integrated Circuits*, 2nd *Ed*, Cambridge, 2004. Sec. 2.3 – 2.5.
2. W. F. Eagan, Practical RF System Design, Wiley, 2003, Chap. 1.

Game Plan for Chap. 2:

1. dB: definition and use
2. Functional blocks in transceivers
3. The outdoor and indoor channel and a way to think about antenna and radar cross section
4. The transceiver block diagram

2.1 dB: definition and use

Decibel is 1/10 of Bel (named after Alexander Bell, although no one uses Bel now). Bel means $\log_{10}$, and therefore, for a unitless power ratio such as SNR (signal-to-noise ratio) in power, dB = $10\log_{10}()$. The logarithmic function has to be applied to a ratio or normalized number without unit. Also, in a signal chain when the transfer function is multiplied, it will become addition in dB. The multiplier 10 makes most of the simple estimation to be in integer only. Some examples are given below.

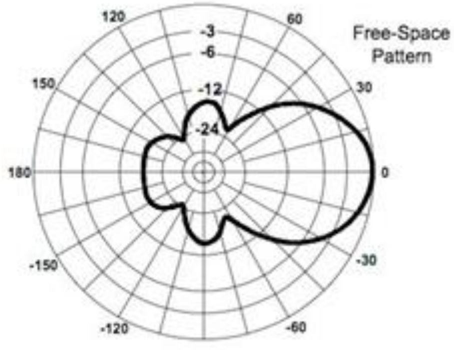
- 2 times in power: $10\log_{10}2 = 3$dB
- Signal power to noise power SNR = 7dB means the signal power if $10^{7/10} = 5$ times larger than noise power.

In RF design, another often used unit is dBm, i.e., power level normalized to 1mW. 1mW is a convenient point as most unit consuming less than 1mW will remain “linear”.

- 0 dBm = $10^{0/10}$ = 1mW
- 1 fW = –120 dBm
- 1 pW = –90 dBm
- 1 W = 30 dBm
- 4 W = 36 dBm
- 1 kW = 60 dBm

Also, if a 3 dBm power goes through an amplifier of 10 dB power gain, it will become 3 dBm + 10 dB = 13 dBm. (Notice how the unit works here). A 30 dBm power in comparison with 0 dBm power is 30 dB or 1,000 times larger.

There are two more popular uses in RF designs:



- dBi in antenna gain: signal power through antennas normalization to a theoretical "isotropic antenna". Notice that the true isotropic antenna, although desirable as often we do not know where the receiving antenna is, does not exist in reality, as there is no spherical solution in the traveling wave. Some asymmetry in the angular coordinates has to exist for far-field radiation solution. For a patch antenna that only emits power to the front 360^o solid angle, the best achievable antenna gain is then: 3 dBi. This means, by power conversation, antennas with high gain have a narrow beam!
- dBc in signal noise: noise power normalized to the "carrier". This is most often used in phase noises and jitter. We will go through those details when we treated phase noise.

In the RF design, it is convenient to describe the voltage signal as sinusoidal continuous waves (CW) of $V_{peak}\cos(\omega t - kz)$ or $V_{peak}\exp(-j(\omega t - kz))$ where V_{peak} is the voltage amplitude. We can thus define the signal power with respect to the peak voltage V_{peak} and root-mean-square (RMS) voltage V_{RMS} as:

$$V_{peak} = \sqrt{2}V_{RMS}\,;\; P = \frac{V_{RMS}^2}{R} = \frac{V_{peak}^2}{2R}\,. \qquad (2.1)$$

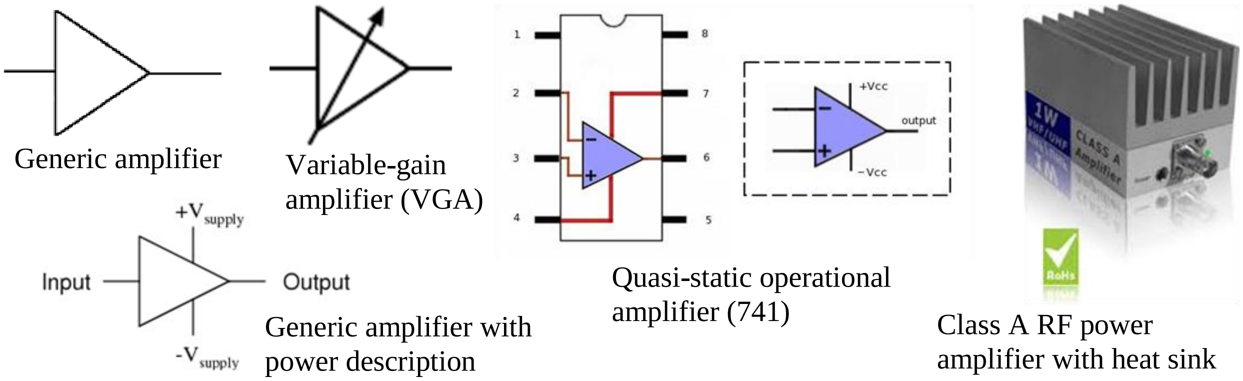
In signal voltage and current, due to the square dependence, the voltage or current ratio becomes $20\log_{10}()$. For example, a voltage gain of 20dB is the voltage magnitude of 10 times larger and the power to increase 100 times. Also, the ratio between two signals at 30dBm and -10dBm is 10^4 times in power (40 dB), or 100 times in voltage (40dB as well).

2.2 Functional blocks in transceivers

Similar to the single LC to discrete transmission line (LC lattice) and distributed transmission line (waveguide), when we describe the signal cascade in RF transceivers, we always have to pay attention to the distinction to quasi-static, continuous wave (CW) and transient. For example, for a step response tracing back and forth on a transmission line, it is a transient description, while the Smith Chart is a CW analysis. When we describe the voltage transfer curve of a digital inverter, it is a "quasi-static" view, which means the "intrinsic" output voltage will always follow the input voltage instantaneously without delay. In radio transceivers, the antenna is most often in CW description, while the microcontroller and data converter (analog-to-digital and digital-to-analog) are most often in quasi-static. For RLC, it can be constructed by lumped elements but used as the impedance match for the CW transmission line. This is why the transmission line described by either distributed and lumped elements is critical in the modeling point of view.

2.2.1 Amplifiers, power amplifiers and low-noise amplifier

Generic amplifiers are represented by a gain factor. Quasi-static amplifiers are often described with a cutoff frequency (f_T), where the output will not only lose the amplification but can be significantly delayed and distorted from the input signal. RF power amplifiers have various classes depending on the configuration and efficiency (Class A, B, AB, C, E, F, etc.). As LC resonator is often used to boost up the output voltage and increase the efficiency, a bandwidth, input and output impedance are additionally required. A general rule is that the higher the power conversion efficiency, the less the bandwidth and linearity. Although we most often considered the power amplifier as unilateral (signal only going from input to output), many power amplifiers have reverse coupling and reflection, which makes the full S or T matrix description necessary (to be introduced later). The amplifier symbols are illustrated in Fig. 2.1.



Generic amplifier

Variable-gain amplifier (VGA)

Generic amplifier with power description

Quasi-static operational amplifier (741)

Class A RF power amplifier with heat sink

Fig. 2.1. Amplifier symbols.

Two additional amplifiers are commonly used in the RF receivers. To increase the radio range, often the receiver needs to be very sensitive to small RF power inputs (common radio: –90dBm; RFID: –70dBm; GPS: –110dBm). No single amplifier can be sufficient. In a chain of amplifiers among the frequency conversion, the first amplifier needs to be a low noise amplifier (LNA) in the RF carrier frequency. LNA usually have gains only around 10 – 20 dB, but very low noise (measured by noise figure, typically around 2 – 6 dB). The bandwidth, and the power level where the gain deviates from constant (called IP3, inter-modulation third-order point), are also described. LNA often has a fixed gain, as most of the design focus will be on the low noise. Remember that the additional noise in a chain of amplifiers will be further amplified by the later stages. Or in another words, the noise in the later-stage amplifiers will be divided by the gain of LNA as seen from the RF input. Therefore, the noise figure in the later-stage amplifier is much less important.

The receiver does not know *a priori* where the transmitter is. To maximize the communication range, the receiver has to be made very sensitive. This presents a problem when the transmitter is very close, i.e., the RF input from the receiving antenna to LNA is too large and cause saturation and overflow of the baseband analog-to-digital (A/D) converter. Often in the intermediate frequency or baseband we will put the variable gain amplifier or even attenuator to mitigate the large RF input saturation. The tuning range can be much easier done in the lower frequency with quasi-static amplifiers when the noise figure is also less critical. The variable gain amplifier is often implemented with a feedback loop with a RF input estimator (can be implemented with a simple RF-to-DC converter), so that the gain control can be automatic. The feedback loop for sure will need stability analysis to avoid oscillation (i.e., any signal in the loop should have loop gain less than 1). The variable gain will also be fed into the microcontroller for an overall scaling from the A/D output.

2.2.2 Filters

RF designs emcompass many frequency conversions for practical antenna and noise reduction. During the conversion, the unwanted frequency (including the image frequency) needs to be filtered out. The frequency components that are kept are called the passband, while the frequency components to have a significant loss are called the stopband. Although RC circuits can create filters, they are not often used due to the loss in the passband. LC circuits with multiple components are more often used in all frequency ranges, especially in the RF frontend. LC bank filter analyses are however rather complicated for the Butterworth (flat passband but less sharp transition to stopband to reduce gain variation) and Chebyshev (sharp transition to stopband with ripples in either pass band or stop band gains for better channel selection) filters. Remember that whatever power consumption in passband or stopband will be heat, which will not only hurt efficiency but need heat sink. A 3dB insertion loss in pass band means half of the signal power will be consumed here, and any power in the stop band will be heat as well!

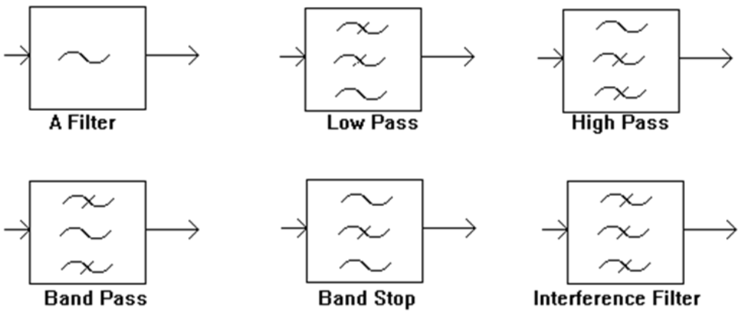


Fig. 2.2. Block Diagram Symbols for Filters

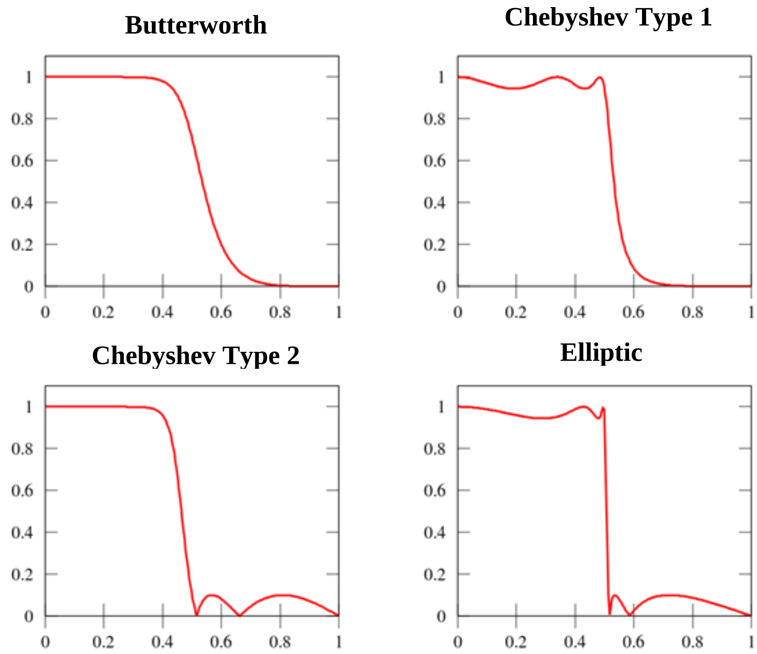


Fig. 2.3. Typical filter transfer functions as a function of normalized frequency.

The Type-I Chebyshev filter has transfer funtion described by:

$$G_n(\omega)=|H_n(j\omega)|=\frac{1}{\sqrt{1+\varepsilon^2 T_n^2\left(\frac{\omega}{\omega_0}\right)}} \qquad (2.2)$$

where ε is the Ripple factor, and T_n is the Chebyshev polynomial of the n-th order. The ripple factor is often expressed in dB by $20\log_{10}\sqrt{1+\varepsilon^2}$ (as a voltage or current conversion). For $\varepsilon = 1$, the ripple factor is 3dB.

When LC tank filters are used (T, π, m-derived, ladder, etc.), as the transfer function is termination dependent, it is important to **match the input and output impedance** to obtain the spectral transfer function. Filters are often bilateral, and we need to be careful for the reverse coupling and the reflection when distributed signals are considered. We will not have time to treat the LC bank filter circuits (in ECE 5790), but will deal with the nonideal issues of filtering in the signal level later. Typical bandpass filter LC tanks are shown in Fig. 2.4 for illustration.

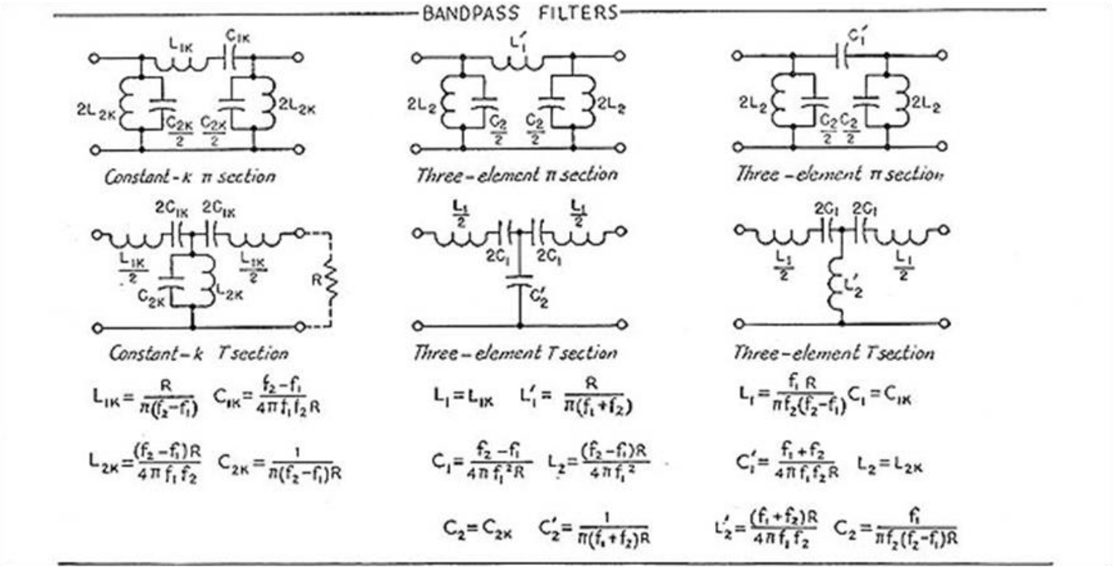


Fig. 2.4. Bandpass filters by LC tank designs.

2.2.3 Frequency mixer

Frequency mixer circuits generate $\omega_1 \pm \omega_2$ frequency components from input of ω_1 and ω_2. This can be achieved with any nonlinear elements when the signal $s_1(\omega_1) \pm s_2(\omega_2)$ goes through a quadratic function (seen as the Taylor expansion of the nonlinear transfer function), or an active multiplier (such as the Gilbert multiplier which comes with a gain as well), or by switching. The basic function is the prosthaphaeresis (product to sum) identity:

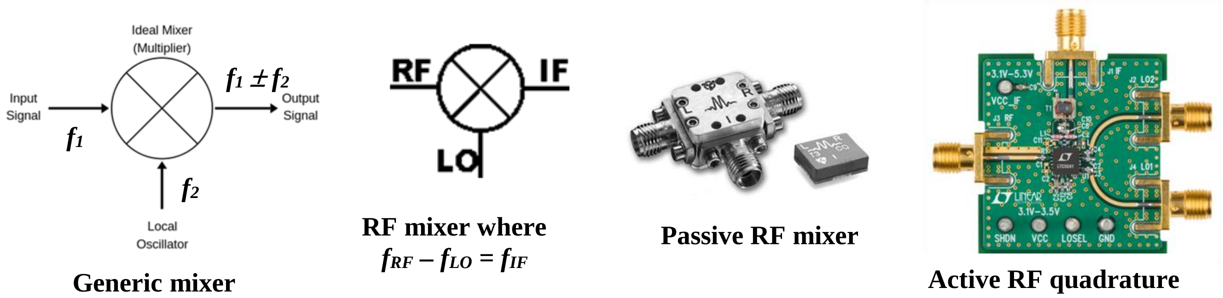
$$\sin(\omega_1 t)\sin(\omega_2 t) = \frac{\cos((\omega_1-\omega_2)t)-\cos((\omega_1+\omega_2)t)}{2} \qquad (2.3)$$

Mixers are typically done in lumped elements (diode or transistors) when the signal level is relatively small, and therefore, mixer description often contains the cutoff frequency f_T and the input and output are taken differentially, with or without impedance match (as the reflection will be small as well). As most mixers use Eq. (2.3), it is often called "multiplier" as well. The most common RF mixer in a superhyterodyne receiver has f_{RF} (radio frequency, i.e., carrier) and f_{LO} (local oscillator from the frequency synthesizer) as the input and f_{IF} (intermediate frequency) as output, by

$$f_{RF} - f_{LO} = f_{IF} \qquad (2.4)$$

Typical mixers are illustrated in Fig. 2.5. Surely we have more than just $\omega_1 \pm \omega_2$ frequency components being generated from the nonlinear mixer function. Other harmonics such as $2\omega_1$, $2\omega_2$, $2\omega_1 - \omega_2$ and $2\omega_2 -$

ω_1 are also important in the signal chain. The "third" harmonics of $2\omega_1 - \omega_2$ and $2\omega_2 - \omega_1$ are often close to the carrier ω_1 band, and can cause serious in-band interference that cannot be easily filtered out.



Generic mixer

RF mixer where $f_{RF} - f_{LO} = f_{IF}$

Passive RF mixer

Active RF quadrature mixer

Fig. 2.5. Mixer schematics and examples

The mixer is also used in the transmitter signal chain. Assume that the baseband data rate can be represented by ω_{base}, the mixing in the transmitter signal chain is:

$$\sin(\omega_{base} t)\sin(\omega_{LO} t) = \frac{\cos((\omega_{LO} - \omega_{base})t) - \cos((\omega_{LO} + \omega_{base})t)}{2} \tag{2.5}$$

Here we did not consider the intermediate frequency of the superheterodyne architecture. Also notice that the baseband signal has more complicated spectrum than represented by a sin() function, which does not carry any data. The ω_{base} is often thought as the baseband data rate or sampling rate in Eq. (2.5). We also notice that the mixing function gives two side bands around ω_{LO}. The implication on spectral efficiency is typically treated in a telecommunication class. We will focus mostly on the handling of image frequency and its resolution in the signal chain.

2.2.4 Amplitude, frequency and quadrature modulation

Monotone carrier waves do not carry data. The carrier frequency is chosen to fit the corresponding antenna and FCC regulation, and its magnitude and phase at the receiver carry only the "travel" information between the transmitter and receiver (such as range or scatterer radar cross section). We need to modulate either the magnitude or the phase or both to represent any data, as in Fig. 2.6.

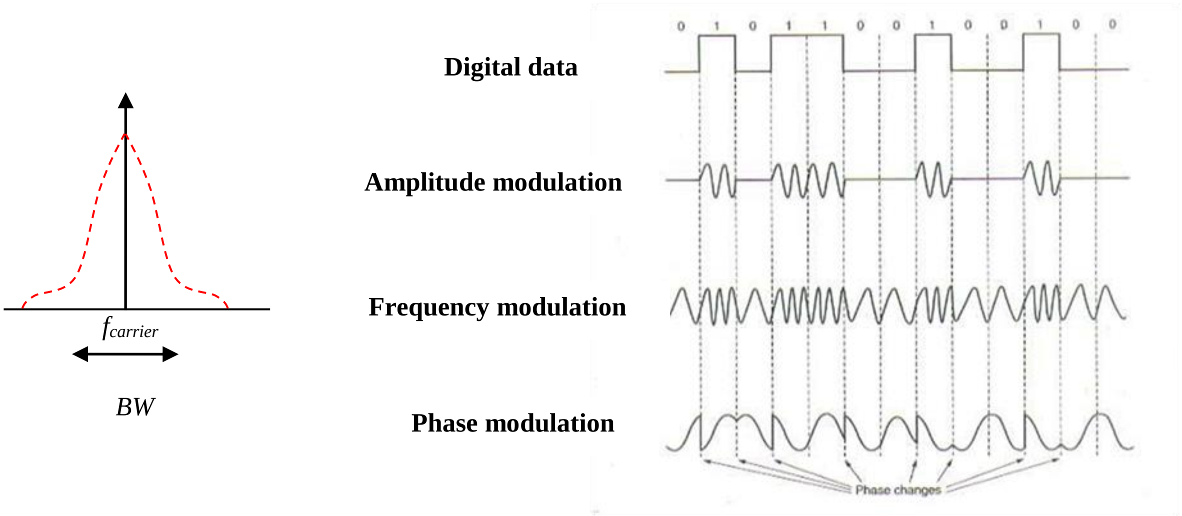


Fig. 2.6. Typical modulation schemes on carriers (exaggerated, as $f_{carrier}$ is often much larger).

The more data are modulated on the carrier, the more the bandwidth *BW* spreading. As the free space is shared, the bandwidth, which relates to the "symbol" rate by Shannon theory, is often much smaller than the $f_{carrier}$. The $f_{carrier} \pm BW$ is called a "channel", which is often just 0.1% – 0.2% of $f_{carrier}$. Within a FCC-regulated band, often 10 – 100 such channels are available to share for frequency hopping (FH) or listen before talk (LBT) schemes. So, the band in FCC regulation typically has bandwidth of 1% – 2% of $f_{carrier}$. Therefore, most of the radio components just need to operate within that band, for example, a frequency modulation (FM) radio. However, radiolocation or radar bands occupy wider bandwidth, as the locating precision is also proportional to BW, typically around 3% - 10%. For more radar discussions, do take ECE 4870. Broader frequency operations within the same radio are possible, but the component design (filters, PA, antennas, etc.) will be much more difficult, or a lot of RF switches need to be employed.

For a typical modulated sinusoidal wave in air $A(t)\cos(\omega_{RF}t + \theta(t))$ where $A(t)$ as the amplitude modulation (AM) and $\theta(t)$ as the phase modulation (PM) change much slower than ω_1, we can use the "quadrature scheme" to harness AM and PM together to achieve a high "symbol" rate. This can be understood from the quadrature modulation function in the receiver signal chain:

$$\begin{aligned} \sin(\omega_{LO}t)\cdot A(t)\cos(\omega_{RF}t+\theta(t)) &= \frac{-\sin((\omega_{RF}-\omega_{LO})t+\theta(t))+\cos((\omega_{RF}+\omega_{LO})t+\theta(t))}{2}A(t) \\ \cos(\omega_{LO}t)\cdot A(t)\cos(\omega_{RF}t+\theta(t)) &= \frac{\cos((\omega_{RF}-\omega_{LO})t+\theta(t))+\cos((\omega_{RF}+\omega_{LO})t+\theta(t))}{2}A(t) \end{aligned} \tag{2.6}$$

After filtering out the high frequency components, we get the quadrature signals *I* (in-phase) and *Q* (90^o phase shift) as:

$$\begin{aligned} I &= A_{mixer}A(t)\cos((\omega_{RF}-\omega_{LO})t+\theta(t)) \\ Q &= -A_{mixer}A(t)\sin((\omega_{RF}-\omega_{LO})t+\theta(t)) \end{aligned} \tag{2.7}$$

We can then evaluate both the AM/PM and the combination with the *I* and *Q* in the baseband (here we use the homodyne frequency conversion without the further complication of the intermediate frequency). For example,

$$A(t) \propto \sqrt{I^2 + Q^2}$$
$$\theta(t) = offset - \arctan\left(\frac{Q}{I}\right) \quad (2.8)$$

Because of the flexibility in the quadrature scheme, and ready/accurate generation of $\sin(\omega_{LO}t)$ and $\cos(\omega_{LO}t)$ in active multipliers, the quadrature multiplier is very popular when the data rate is high. As A(t) and θ(t) can have multiple levels, e.x., $A(t)$ has 4 distinctive levels of 0, 1, 2 and 3 and $\theta(t)$ has another four levels of 0^o, 90^o, 180^o and 270^o (depending on implementation this can be 0^o, 45^o, 90^o and 135^o instead), we can in total implement 4-bit per "symbol" (16 states by the combination of amplitude and phase) in the quadrature modulation.

Quadrature modulation also improves linearity from the signal splitting, which we will cover more when we introduce different branch and loop architecture. We have not discussed the frequency modulation (FM) as shown in Fig. 2.6. For a small frequency deviation f_Δ to the carrier frequency, the modulated signal can be modeled as:

$$y(t) = A\cos\left(2\pi\int_0^t f(\tau)d\tau\right) = A\cos\left(2\pi\int_0^t \left(f_{LO} + f_\Delta m(\tau)\right)d\tau\right) = A\cos\left(2\pi f_{LO}t + 2\pi f_\Delta \int_0^t m(\tau)d\tau\right) \quad (2.9)$$

FM is similar to the large sinusoidal phase modulation, known as the Carson's rule.

2.2.5 Data converter

Data converter includes the function of analog-to-digital (A/D) and digital-to-analog (D/A). As the RF signal in air is analog, A/D is used in the end of receiver and D/A is used in the beginning of the transmitter. Data converter is often the interface between digital backend and the RF frontend in a transceiver, but fast digital electronics today have smeared the boundary. Many of the RF frontend control can be done digitally, and should be done digitally, as digital information is regenerative[1] (the noise can be cleaned during digital regeneration) and error correctable, while the analog information is forever locked to the signal-to-noise ratio (SNR)! As the data converter is in the "baseband" for data, the frequency is usually low. The frequency range where the data can be converted correctly is called the bandwidth. Together with the tolerance of SNR, the bit length and the bandwidth are the sufficient description for data converters. A 20kHz sampling is often considered reasonable, and many microcontrollers give you the A/D and D/A 10-bit converters (about 1,000 discretized levels) in that sampling range already. Implementation of the fast and efficient data converter circuits deserves its own course, and will not be included in 4880.

2.2.6 Circulator

An RF circulator is a "non-reciprocal three-port device where the RF signal entering any port is ONLY transmitted to the next port in rotation. The circulator can be passive or active. The symbol and the transfer function are shown in Fig. 2.7. The loss for the pass signal is called the "insertion loss" (as in many other passband component), and the ratio between the pass signal and the stop signal (for below said g_{12}/g_{13}) is called the "isolation", which is typically about 40 – 60dB. Circulator is mostly used in the "duplex" circuits to share the antenna between the transmitter and the receiver without too much self

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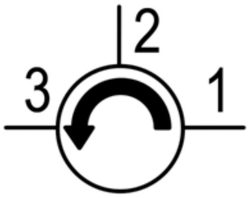
[1] "Regenerative" has also been used as a radio receiver architecture, where the receiver will gradually saturate to oscillation if not reset periodically. Digital signals are interpreted as enhancement or inhibition of the time to oscillate.

jamming (the direct path from the transmitter to the receiver is blocked by the isolation). Half duplex (like the old walkie-talkie or ham radio) is referred to time division of transmitting and receiving with further RF switches added, and full duplex is referred to frequency or code division of transmitting and receiving (like your cell phone). Circulators are added to further improve isolation between the transmitter and the receiver.



$$S = \begin{pmatrix} 0 & 0 & 1 \\ 1 & 0 & 0 \\ 0 & 1 & 0 \end{pmatrix}$$

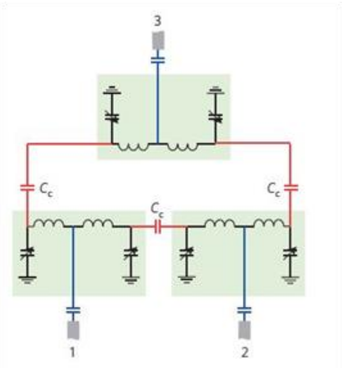
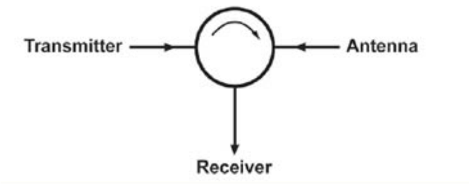


Fig. 2.7. Circulator symbol, ideal function and module.

Notice that if your radio has a small form factor, two close-by transmitting and receiving antennas will NOT replace the circulator. When two antennas are close to each other (within $\lambda/2\pi$), they will detune each other, and function like ONE antenna. When they are separated by more than $\lambda/2\pi$, traveling wave can be produced from one antenna to the other, the self interference can still be very strong.

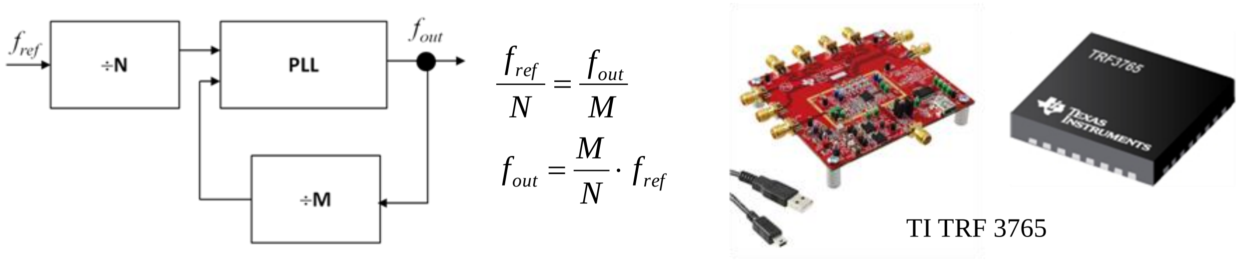
Exercise:

For the circulator before the antenna, list all of the strong leaky paths to the receiver chain.



2.2.7 Frequency synthesizer: local oscillator, frequency reference

The local oscillator (LO) is of ultimate importance to the radio transceiver in performing frequency conversion. As the transmitter-to-receiver link is at two locations and hence two different manufacturing (and possibly two different temperatures), the modulation and demodulation will ONLY work accurately if the two LOs are very much the same (a process called synchronization). The amplitude and phase noises of LO are also critical, as those are directly added to the signal chain as well. The frequency synthesizer becomes even more important for the “cognitive radio” (i.e., the channel select is done after evaluation of the present situation) in a multiple-input multiple-output (MIMO) network, as LO may need to change frequency even faster than the conventional frequency hopping scheme. Classical frequency synthesizer is made by the rational divider-PLL feedback loop as shown in Fig. 2.8. f_{out} can inherit the high quality in f_{ref} for this rational synthesizer. The divider can be implemented digitally with counter circuits or special parametric amplifiers. Changing f_{out} frequency by digital control of N and M often comes with a delay much more than 1,000 cycles of f_{out} (for 1GHz, this is about 1μs), which needs to be taken into consideration during channel selection by mixers.



$$\frac{f_{ref}}{N} = \frac{f_{out}}{M}$$

$$f_{out} = \frac{M}{N} \cdot f_{ref}$$

Fig. 2.8. Classical rational frequency synthesizer.

The reference frequency f_{ref} is often from a crystal oscillator where the frequency is accurate and independent of temperature. At the same time the phase noise is small. The crystal oscillator is an electromechanical device based on the piezoelectric resonance of a specially cut quartz as shown in Fig. 2.9. The operation can be reasonable modeled by the van Dyke resonator circuits also as shown. The quality factor Q is often above 1,000 and the temperature coefficient is typically very small.

$$f = f_0\left(1 - 0.04\,ppm(T - T_0)^2\right) \qquad (2.10)$$

The quartz oscillator was inserted to radio broadcast in 1920s. Before that, a typical LC resonator was used and suffered significantly more “drift” and channel interference. An old AM radio was given 10kHz bandwidth (with human hearing between 20 to 20kHz, you can see AM music is not great for music, but reasonable for speech of 1kHz to 4kHz) in a carrier around 1MHz (medium wave, while short wave radio covers around 10MHz). A drift of 3kHz in LC resonator due to the component or temperature can cause the neighboring channel to interfere each other. After the introduction of quartz oscillator, this is much more relieved, as the drift is often controlled below 1ppm.

The quartz oscillator is also used to generate the “clock signal” in digital watches and computer chips today. As this is an electromechanical component, many experiences had been passed down to a “microelectromechanical” (MEMS) resonator, which has worked towards the replacement of the quartz oscillator as an chip-integrated version. MEMS resonators, including small vibrating structures, surface and bulk acoustic modes, piezoelectric materials, etc., had been seriously investigated in the last 20 years.

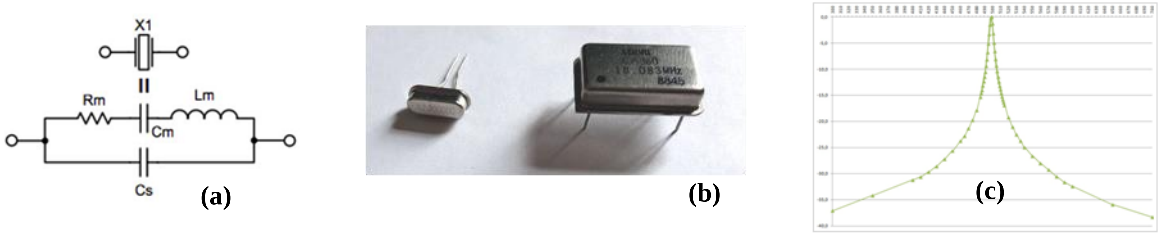


Fig. 2.9. Crystal oscillators based on piezoelectric quartz crystal vibration: (a) Circuit symbols and van Dyke equivalent circuits; (b) Hermetic packaging to reduce the influence from ambient air; (c) Typical resonator response with $Q > 1{,}000$.

2.3 Outdoor and indoor channel modeling

2.3.1 Free Space Loss

Free space loss of wireless signals can be modeled by the Frii’s far-field equation:

$$P_R = P_T \left(\frac{\lambda}{4\pi r} \right)^2 \Psi_T \Psi_R \qquad (2.11)$$

where P_T and P_R are the signal strength in the transmitter and receiver in dBm, and Ψ_T and Ψ_R are the transmitter and the receiver antenna gains in dBi, respectively. λ is the wavelength of the traveling wave and r is the distance. Frii's equation is simply power conversation considering the source as a point, the receiver as an area on the sphere and no additional loss in the far-field propagation ignoring attenuation such as in vacuum or space. It is not really "loss", but "spreading". This is where the $4\pi r^2$ term enters Eq. (2.11). The wavelength λ dependence is somewhat artificial, but practical and convenient. The receiving antenna gain Ψ_R is normalized to the imaginary half-wavelength isotropic antenna (dBi), i.e., we assume the "radar cross section" (RCS) of the receiving antenna is proportional to the wavelength. Hence, small wavelength will use a small-area antenna as far as Eq. (2.11) is concerned, which captures less power in the $4\pi r^2$ spherical surface. If you happen to use a larger antenna[2] in the high-frequency range, such increase in the antenna power collection will be absorbed to the receiver antenna gain, similar to the beamforming direction effects. It is convenient to remember the free-space loss in frequency close to the ISM (industrial, scientific and medical) unlicensed bands which are NOT intended for telecommunication, i.e., micro-broadcasting, cordless phones, Bluetooth, etc. (13.56MHz, 26.96MHz, 40.6MHz, 433.0MHz, 902MHz, 2.40GHz, 5.725GHz, and 24.0GHz):

Table 2.1. Free space loss for typical frequencies and ranges.

Free Space Path Loss at P_S = 0dBm, $\boldsymbol{\Psi_T = \Psi_R}$ = 0dBi	**at 1m**	**at 10m**	**at 100m**	**at 1km**
1.2GHz	–34dB	–54dB	–74dB	–94dB
2.4GHz	–40dB	–60dB	–80dB	–100dB
5.0GHz	–46dB	–66dB	–86dB	–106dB
50GHz	–66dB	–86dB	–106dB	–126dB

The free space loss is the minimal loss for direct line-of-sight (LoS) transmission (work well in space), but would need two important corrections: attenuation in air/ionsphere and multi-paths for cities and indoors.

2.3.2 Attenuation and reflection in air

Frii's free space loss assumes far-field ($r > \lambda/2\pi$) wave propagation and hence the energy travels on the expanding spherical surface. For air, there is addition dielectric loss for air with variable H_2O content to be included:

$$\vec{E} = \hat{r}\, E_r(r - ct) = \hat{r} E_o \exp\left(j\omega t - j\left(k - j\frac{\alpha}{2} \right) r \right) = \hat{r} E_o e^{-\frac{\alpha r}{2}} \exp\left(j(\omega t - kr) \right) \qquad (2.12)$$

where α is the power attenuation constant in cm^{-1}. We can see that attenuation appears as a prefactor and is more typically expressed in dB/m or dB/km. In the frequency range of 1M – 10GHz, attenuation can

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[2] Antenna smaller than quarter wavelength will drop the antenna gain very fast, as the current variation in the Hertzian dipole cannot even finish peak to valley changes. However, larger antenna does not increase the gain as fast! Detailed geometry needs to take the full effect of wave periodicity which limits the antenna bandwidth.

often be ignored except in heavy rain. For higher frequency, attenuation is a function of the temperature, pressure and humidity. Typical weather-dependent attenuation is shown in Fig. 2.10[3]. Although air molecules are very sparse in comparison to RF frequencies below 40GHz, mist/cloud as small water droplets can severely absorb RF energy above 40 GHz, especially in the proximity of the resonance frequency of H_2O and O_2.

Unlike light, RF signals can penetrate mild solid obstruction as well, with additional loss from reflection and from dielectric loss. It is not strange that a wall-penetrating radar is in active service today! Typical additional loss for one-way penetration is listed in Table 2.2. The rule of thumb is: anything thinner than $\lambda/2\pi$ will have some penetration, as the solution does not depend on the traveling wave decay.

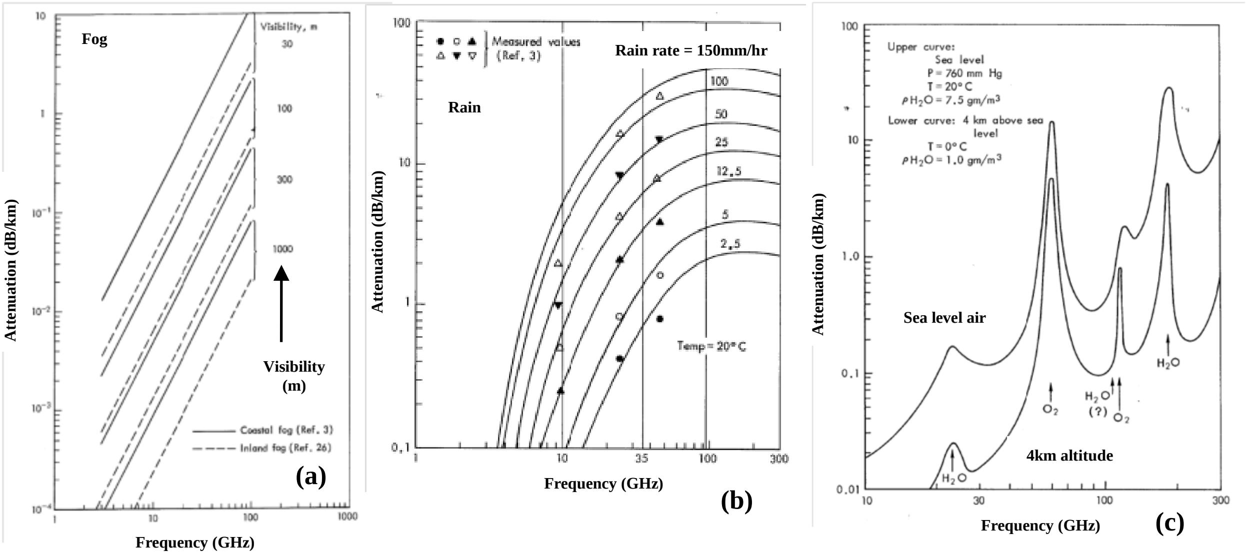

Fig. 2.10. Attenuation (dB/km) in various weather conditions: (a) In fog characterized by visibility; (b) In rain characterized by rain rate; (c) Horizontal propagation attenuation with O_2 and H_2O absorption bands.

Table 2.2 Loss of RF signals around 1GHz for indoor partitions

Partition type	Typical partition loss in dB
Cloth/drape	1.4
Double plasterboard wall	3.4
Aluminum foil (20 lbs sheet)	3.9
Concrete wall	13
Aluminum siding	20

2.3.3 Multipath: reflection and diffraction

The previous free-space loss and attenuation assume line-of-sight (LoS), which is the minimal loss. You probably have noticed your satellite GPS and radios do not work well when you lost the LoS path with the southern sky! For cities, forest and indoors, when LoS between the transmitter and the receiver is blocked, multi-path reflection and wave diffraction according to the Huygen's principle can still give sufficient signal power to the receiver with more free-space loss. Due to the importance of multi-path effects, many models have been proposed, including the two-ray and ten-ray models in Fig. 2.11[4]. Other

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[3] C. C. Chen, "Attenuation of electronmagnetic radiation by haze, fog, clouds and rain", *US Air Force RAND Report*, 1975.

[4] A. Goldsmith, *Wireless Communications*, Chap. 2, Stanford University, 2004.

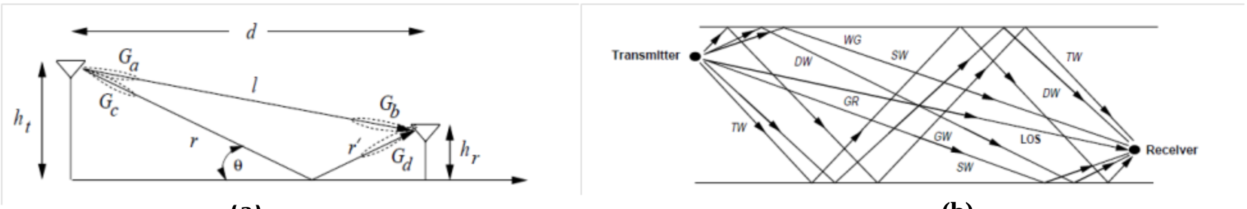
popular models include Fresnel knife edge diffraction, Okumura's and Hata, Bertoni, piecewise, etc. A simplified model for the path loss is most convenient for RF engineers to make estimation calibrated with a single measurement:

$$P_R = P_T K\left(\frac{d_0}{d}\right)^{\gamma}\text{; or}$$

$$P_R(dBm) = P_T(dBm) + K(dB) - 10\gamma\log_{10}\left(\frac{d}{d_0}\right) \qquad (2.13)$$

The free-space loss on wavelength and antenna gains are absorbed in K and d_0. The calibration step is to measure $K = P_R - P_T$ when $d = d_0$.

The multi-path effect will be represented by the parameter γ, which is often determined by least-square fit of the multiple measurements at various d in Eq. (2.13). For the two-ray model in Fig. 2.11, we will have $\gamma \cong 4$ because of the one reflection. It can be understood that the reflection strength will depend on $1/d^2$ and then start another $1/d^2$ free space travel. For multi-path experiments, we list the typical values of γ in Table 2.3.



(a) **(b)**

Fig. 2.11. Multi-path models for free-space loss estimation: (a) Two-ray model; (b) Ten-ray model.

Table 2.3 Decay coefficient γ in Eq. (2.13)

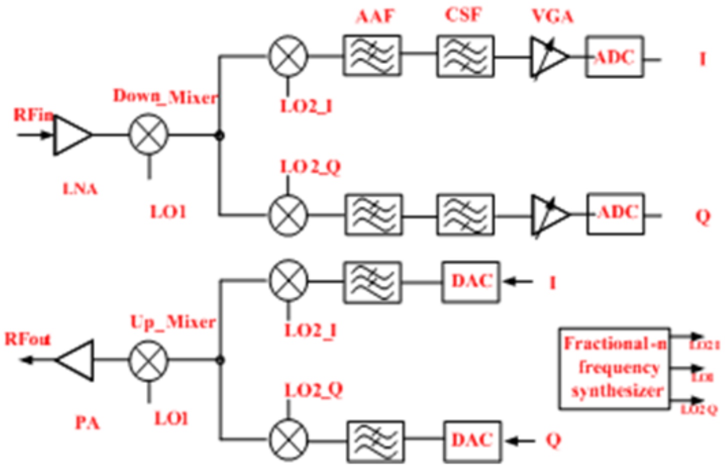
Environment	γ range
Urban macrocells	3.7 – 6.5
Urban microcells	2.7 – 3.5
Indoor office building (same floor)	1.6 – 3.5
Indoor office building (multiple floors)	2 – 6
Large store or warehouse	1.8 – 2.2
Factory floor	1.6 – 3.3
Home (wood)	2.9 – 3.1

We can see the larger the average number of scattering in the multi-path, the larger the parameter γ, and the faster the free-space loss decays. γ can be smaller than 2 when the multi-path effect actually combines with the LoS transmission, such as in indoor office building. Within the outdoor environment, the easiest way to have γ close to 2 is to put one of the antenna very high in position. Now you can see why we would like to build high towers for broadcasting radio and TV! Most of the base stations for your cellular network are implemented on top of road lamp post!

2.4 Ideal transceiver block diagram

We will use what we understand about the component to trace the signal flow in a radio system in the most simplified way. We will leave the complications of gain reflection, noise addition, nonlinearity and intermodulation to later treatment with details. We will define system specification of maximum range of operation, bandwidth, link budget and dynamic range of the receiver.

A typical superheterodyne transceiver is shown in Fig. 2.12. We cannot effectively explain the exact advantages of the use of the intermediate frequency yet (noise related), and will temporarily ignore that part. We will restrict our discussion to a point-to-point radio link in Fig. 2.13, where the other transceiver with the identical transceiver structure knows exactly what channel to transmit, and the LO on two transceivers can be considered "synchronized" (i.e., they are on the same clock). We will temporarily ignore any design techniques towards synchronization, especially in a multiple-transmitter scenario. We will assume the crystal oscillator gives out exact frequency and the rational frequency synthesizer has little quantization error. We use the link between the transmitter of Radio 1 and the receiver of Radio 2 as an example.



- LNA: low-noise amplifier
- I, Q: quadrature signals
- Down_mixer: down-conversion mixer
- LO: local oscillator
- AAF: anti-aliasing filter (low-pass)
- CSF: channel select filter (band-pass)
- VGA: variable gain amplifier
- ADC: analog-to-digital converter
- DAC: digital-to-analog converter
- Up_mixer: up-conversion mixer
- PA: power amplifier

Fig. 2.12. An example of the superheterodyne transceiver.

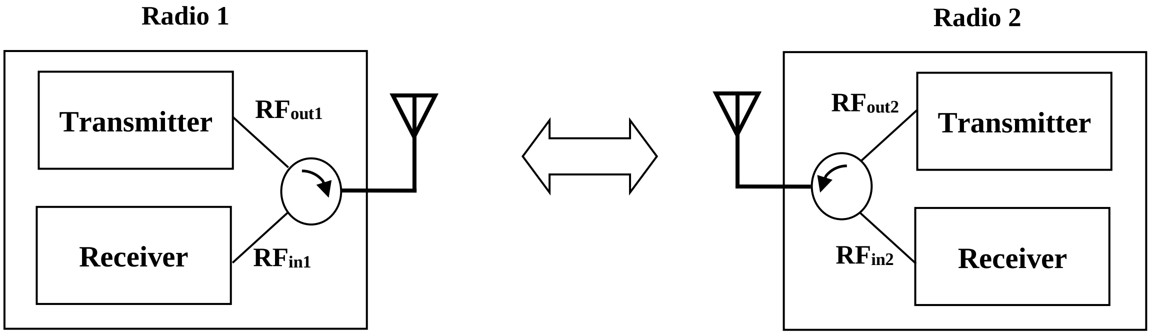


Fig. 2.13. Point-to-point duplex radio link.

1. First, we decide to use the **2.4GHz band** as the radio link, and the desirable operation distance is 1,000m LoS.
2. We decide that we cannot assume any orientation of the two radio units, and semi-isotropic whip antennas of −3dBi will be used.
3. The ideal free space loss is at −100 dB estimated from Table 2.1. Notice that due to the multipath and partition loss, the indoor range is typically only about 10 – 100m.
4. The **link budget** is then: LB = −100 dB – 3dB – 3dB = −106dB.

5. We look up FCC for this band, and decide to go for the approved **Zigbee** protocol (IEEE 802.15.4), which allows maximum RF_{out1} = 20 dBm transmitting power.[5] The link budget then dictates that the **receiver sensitivity** needs to be RF_{in2} = 20 dBm – 106 dB = –86 dBm.
6. If the circulator in Radio 1 provides 60dB isolation, RF_{in1} needs to tolerate –40 dBm in-band, out-of-channel **self interference**, which will enter part of the mixer of $LO1_{transmitter}$ and $LO1_{receiver}$ in Radio 1 mixer design.
7. Considering the receiver sensitivity at –86 dBm, the mixer will provide –40 dBm – (–86 dBm) = 46 dB out-of-channel rejection. If this is not achievable, we will need further RF filters to help channel selection.
8. Also, if we decide the two radios need to function at 1m distance, $RF_{in2,max}$ will be at 20 dBm – 40 dB – 3 dBi – 3 dBi = –26 dBm. Assume this is the strongest signal present (in multiple-transmitter scenario, this can be even worse), the receiver needs a **dynamic range** of at least –26 dBm – (–86 dBm) = 60 dB. Often the higher signal power needs to be tolerated for protection against jammers or severe noise. Therefore, the dynamic range of the receiving strength is often much higher than 60 dB.
9. Now we can look at inside the transceiver. For the receiver, –86 dBm on a 50Ω antenna corresponds to a sinusoidal wave of $P = V_p^2/2Z$ at V_p = 16μV. The gain is provided by LNA, the two mixers and the variable gain amplifier. For a series of amplifiers, the noise is most important for the first stage, as it will be amplified by the following stages. Or alternatively, if we view all noises from RF_{in}, then the noise in the latter stage will be divided by the gain of the previous stages. Therefore, LNA is most critical for its low noise figure with reasonable gains. The function of variable gain control is often left for latter stages in much lower frequency, where the amplifier works in the quasi-static mode, and the impedance matching is much less critical.
10. The low-pass anti-aliasing filter will rid of any RF components, while the channel select bandpass filters perform further filtering in multiple-channel combination or in assisting the data sampling in the data converter.
11. The quadrature modulation is performed when the signal has been demodulated from the carrier, so no special transmission line splitting is needed.

For sure, realistic wireless network most often needs full duplex and multiple access (MA), such as your cell phone and Wi-Fi. Broadcasting of radio and TV, and reception of beacon signals from base stations are notable exception, where only receiver or transmitter is needed. There are in general three MA protocols: time division, frequency division and code division (TDMA, FDMA and CDMA). TDMA is used in half duplex, or during the polling stage when unknown number of users are involved. FDMA is present in many full duplex transceivers, when the transmitter and the receiver are on different "channels" within the same band. Channel selection has to be performed by additional filters before or after the mixer, as the antenna and circulators have to allow all channels in the given band.

Exercise:

For radar or RFID readers listening to the echo of the reflected signal as shown below, define the link budget and dynamic range of the reader. Also, explain why only one LO is sufficient (this is called the coherent receiver, which is true for many radar and RFID situation when "echoing" is used)?

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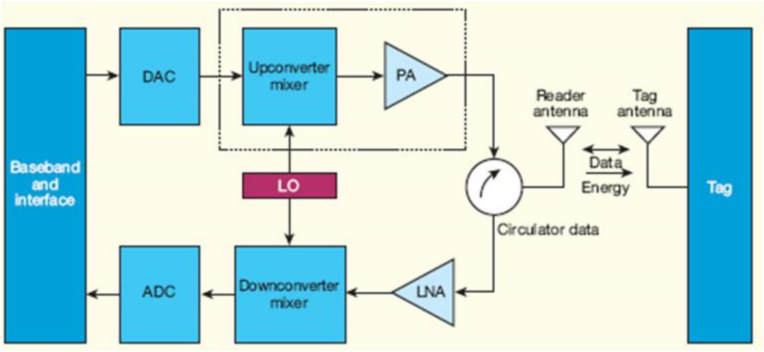
[5] Zigbee also governs the channel bandwidth (2MHz), channel separation (5MHz) and frequency hopping among the 16 channels. The data rate is at 250 kb/s with offset quadrature phase-shift keying (OQPSK 2 bits per symbol). Zigbee has relatively low bit rate for its bandwidth, optimizing for the operation under severe noises.


Baseband and interface
DAC
Upconverter mixer
PA
LO
ADC
Downconverter mixer
LNA
Reader antenna
Tag antenna
Data
Energy
Circulator data
Tag