# Basic concepts

## Big picture

How have we been able to make cell phones more affordable while their functionality has exploded (cellular, WiFi, Bluetooth, GPS, computing, storage, digital camera, user-friendly interface)?

* Integration: how much functionality can be placed on a single chip (or how few components are left off-chip)
* Integration is due to (1) scaling of VLSI processes – CMOS technology – (2) innovations in RF architectures, circuits, devices.

These disciplines are all required, to some degree, for an RF designer:

* Communication theory
* Random signals
* Transceiver architectures
* IC design
* CAD tools
* Wireless standards
* Multiple access
* Signal propagation
* Microwave theory

RF design hexagon:

* Noise
* Power
* Frequency
* Gain
* Supply voltage
* Linearity
* … back to noise

Each metric trades off with its two adjacent metrics. For example, to lower the noise of a front-end amplifier, we must consume more power or sacrifice linearity.

Generic RF transceiver architecture:

* Receiver: antenna 🡪 LNA 🡪 downconverter (driven by oscillator, generated by frequency synthesizer) 🡪 ADC 🡪 digital baseband processor
* Transmitter: digital baseband processor 🡪 DAC 🡪 upconverter 🡪 PA 🡪 antenna

## Units in RF design

Voltage gain:

Power gain:

**Voltage and power gain are equal in dB if and only if the input and output voltages appear across equal impedances.** For example, the gain of an amplifier with an input resistance of and driving a load of is

and are rms values.

Powers are expressed in dBm:

For example, if we deliver a power of 0dBm across a 50Ohm load for a sinusoidal signal, what is the peak-to-peak voltage swing?

Another example, a GSM receiver senses a narrowband modulated signal having a level of -100dBm. If the front-end amplifier has a voltage gain of 15dB, what is the peak-to-peak voltage swing at the amplifier output?

We assumed that

* The input impedance of the front-end amplifier is 50Ohm
* A narrowband signal has roughly the same peak-to-average power relationship as a sinusoid

In most integrated RF systems, we prefer voltage quantities to power quantities since

* Input and output impedances of cascade stages may be unequal, so voltage gain and power gain are not equal
* Impedances may be largely capacitive or inductive, in which case there is no “real” (active) power

However, we still sometimes use dBm at interfaces that do not necessarily entail power transfer. If we drive a purely-capacitive load, the delivered average power is zero, but we can still calculate dBm as if we were driving a 50Ohm load with our voltage signal.

## Time variance

A system is linear if and only if it satisfies the principle of superposition.

Systems with nonzero initial conditions or DC offsets are technically nonlinear, but we often relax the rule to accommodate these two effects (in this case, the system is incrementally linear).

A system is time invariant if a time-shift in the input causes the same time-shift in the output.

A system that changes with time is time variant.

Take the ideal switch for example. Let drive the control s.t. the switch is on if , and let drive the input.

If we look at as the input and as part of the system, then the system is both nonlinear and time variant. The output is independent of the amplitude of , and the system varies over time with .

If we look at as the input and as part of the system, then the system is linear but time variant. The output scales with , but the system varies over time with . In this case, the input-output relationship is

is a square wave toggling between 0 and 1 with frequency .

The output spectrum consists of copies of at

**A linear system can generate frequency components that don’t exist in the input signal if the system is time variant.**

## Nonlinearity

A system is “memoryless” or “static” if its output does not depend on past values of the input and/or output.

A memoryless linear system is given by

A memoryless nonlinear system is given by

are functions of time if the system is time variant.

When for even , the nonlinear system has odd symmetry, which means that is an odd function of , i.e.

This kind of system is also called balanced. One example: a differential pair.

A system is “memory” or “dynamic” if its output depends on past values of input and/or output.

An LTI dynamic system is represented by

This is the convolution integral.

If a dynamic system is linear but time variant, its impulse response depends on the time origin. Then

If a system is both nonlinear and dynamic (what about time variant?), then its impulse response can be approximated by a Volterra series.

**Example: square-law MOS transistor operating in saturation**

For a differential pair, the transfer curve is

Factoring out from the square root and assuming , and using , we get

The first term is the small signal gain . Due to symmetry, there is no even-order nonlinearity. Square-law devices generate a third-order term.

## Effects of memoryless nonlinearity

Model:

is the small signal gain of the system.

We will analyze the effect of nonlinearity on sinusoidal inputs.

### Harmonic distortion (signal only)

Let the input be a single real sinusoid.

Even-order nonlinearity introduces the DC offset. Ideally, even-order nonlinearity vanishes in balanced circuits, but random mismatches corrupt the symmetry, yielding finite even-order harmonics.

th-order harmonic amplitude is proportional to .

**RF harmonics are typically less critical because they fall way outside the frequencies of interest.**

**However, you need to be careful if mixing occurs.**

For example, let’s say are the two inputs to an analog multiplier (mixer). The ideal multiplier is modeled as

is a constant.

Let .

The ideal output has frequency components .

What happens if experiences third-order nonlinearity at the input?

Then the output will have spurious frequency components . Let . Let’s say the desired output frequency component is . , which can fall inband – this is a problem.

### Gain compression (signal + 1 interferer)

Nonlinearity compresses small signal gain, as seen in the fundamental component

Where and have opposite signs for compressive behavior.

The 1dB compression point – the point at which gain drops by 1dB – is one standard way of characterizing nonlinearity.

In general,

Note that is the peak value of the sinusoidal input.

**Often, gain compression is not dominated by your desired signal but by a large interferer.**

Let

If you multiply out , you will get a term .

Then the fundamental is

Since ,

As interferer amplitude increases, gain decreases and can even become 0.

### Cross modulation (signal + 1 AM interferer)

What if the interferer has amplitude modulation, i.e. ?

is a constant. Then

The desired signal suffers from amplitude modulation at and .

Cross modulation commonly occurs in systems that must simultaneously process multiple independent channels.

**In a memoryless nonlinear system, cross modulation does not occur if the interferer has phase modulation, e.g.**.

### Intermodulation (signal + 2 interferers)

The two interferers mix inside the nonlinear system.

If are close in frequency, then the 3rd-order products at are close to .

Let’s say we have a signal at , and it happens that . The IM3 product falls directly on the desired signal and corrupts it.

**For IM3, we typically model narrowband signals by condensing them into unmodulated tones.**

**Even if gain is not significantly compressed, IM3 products may still severely corrupt the desired signal.**

Example: we have a LNA with a gain of 10 and input impedance of 50Ohms. The LNA senses a desired signal level of -80dBm @ 2.41GHz and two interferers of equal power at 2.42 and 2.43GHz. Assume the LNA drives a 50Ohm load.

What value of yields a 1dB compression point of -30dBm? -30dBm is average power.

If each interferer is 10dB below , what is the corruption of the desired signal?

The IM3 at the LNA output is given by

The signal at the LNA output is , so the IM product is as large as the signal even though the LNA does not experience significant compression.

#### IP3 and two-tone test

The two-tone test is very useful for characterizing the nonlinearity of the system since the frequency difference can be made arbitrarily small to ensure the IM3 product lands inband (unlike HD).

We apply two sinusoids of equal amplitude, representing the interferers, and normalize the IM3 amplitude to the fundamental amplitude at the system output. This gives us relative IM:

When increases by 6dB (2x), IM3 increases by 18dB and relative IM3 increases by 12dB.

IP3 is a measure of IM3 nonlinearity that is independent of . The idea is that as increases, at a (theoretical) point, relative IM3 becomes 0dBc. In terms of amplitude, fundamental increases 20dB/decade, while IM3 increases 60dB/decade.

This point is

is peak input amplitude. At this (theoretical) point, the output amplitude of one IM3 tone is equal to the output amplitude of one fundamental tone.

Note: is a theoretical quantity – it cannot be directly measured. It is 9.6dB higher than , which means we cannot say that the output fundamental is because of gain compression. Furthermore, may be higher than the supply voltage. When measuring , we must ensure we measure in the region where 1dB increase in results in 3dB increase in IM3.

Example: LNA senses -80dBm signal at 2.41GHz and two -20dBm interferers at 2.42 and 2.43GHz. What is the value of IP3 that results in IM3 being 20dB below the desired signal? Assume 50Ohm interfaces.

Typically, we measure IIP3 using extrapolation. We measure IM3 at different input amplitudes and extrapolate out to the point where IM3 and fundamental intersect.

However, there is a shortcut that provides an estimate (true IIP3 may differ if there is dynamic nonlinearity).

Let be the power of one input tone. is the power of one output fundamental tone, and is the power of one output IM product.

Theoretically, every 1dB backoff from equals 2dB backoff in relative IM3.

This equation requires only one measurement.

It’s often useful to calculate the input-referred IM3 level when you know .

### Cascaded IM3

Let’s say you have two cascaded nonlinear stages

Expanding and keeping only first and third-order terms,

Then the IIP3 of the cascade is

At this point, let’s examine how the IM3 are generated.

1. IM3 generated in first stage
2. IM3 generated in second stage
3. First stage generates HD2, then second stage mixes fundamental and HD2 to form IM3.

In real systems, this last term will be very small because (1) 2nd-order nonlinearity is suppressed by odd symmetry (2) HD2 is suppressed in a narrowband system.

Then we can write

That is, referred to the input, is reduced by the gain of the first stage.

For stages,

This means that IIP3 becomes more important as you progress along a chain of amplifiers because the IIP3 of a given stage, referred to the overall input, is scaled down by the amplification of the previous stages.

In this analysis, we assumed that the IM3 products of each stage add in-phase (that is, there is no destructive cancellation). In a real system that has memory, there will be phase shifts that may result in finite cancellation. However, since are similar frequencies, we expect them to experience similar phase shifts.

**Example:** LNA with IIP3 of -10dBm and a gain of 20dB is followed by a mixer with IIP3 of +4dBm. Which stage limits the overall IIP3?

The contribution of the second stage is

The second stage limits IIP3.

### AM/PM conversion (APC) – TBD include first-order example

Phase shift depends on signal amplitude.

APC doesn’t occur in LTI systems because in LTI systems, phase shift is a function only of frequency.

APC doesn’t occur in a memoryless nonlinear system because phase shift is 0.

**Therefore, APC occurs only in a dynamic, nonlinear system.**

## Noise

### Noise as a random process

Noise is random, which means we cannot predict the instantaneous value. However, its statistics are quantifiable.

For example, if is the noise waveform, average noise power is

This is in analogy with periodic power signals. This is energy over time. To measure , depends on the frequencies of interest. must be long enough to capture several cycles of the lowest frequency component. In practice, we make a guess for and measure . Keep increasing until converges.

### Noise spectrum

In time domain, the only measure of noise we can get is average power.

Knowing the noise spectrum – average power for each frequency component – provides much more information.

Finding the spectrum or PSD of any signal (or noise) is useful.

Conceptually, if you want to know the power at frequency , you pass the signal through a 1Hz-bandwidth brickwall filter at and measure average power.

Do this for all frequency components and you get PSD of , , which has units of or . is power in a 1Hz bandwidth for all frequencies. Conceptually, this is how spectrum analyzers work.

In signals and systems, the PSD is defined as the Fourier transform of the autocorrelation of a signal. These two views are equivalent.

(Parseval’s theorem) The total area under represents the average power of :

Note: this definition is for a single-sided PSD and assumes that is real, because if is real, then PSD is symmetric (PSD is always real). If PSD is symmetric, then it can either be defined as one-sided or two-sided. The two-sided PSD is the same as one-sided except scaled by ½ and symmetric around DC.

**Example: a resistor of value generates a noise voltage whose one-sided PSD is given by**

is absolute temperature (which is in Kelvin). is the Boltzmann constant and is equal to 1.38e-23 .

This flat PSD is called white noise because, like white light, it contains all frequencies with equal power levels.

The area under the PSD is infinite, implying that noise power is infinite. In reality,

If you were to integover all frequency, you would get infinite power. In reality, begins to fall off at , i.e. thermal noise is not truly white.

The dimension of is .

In fact, if we denote the noise voltage by , we can write

denotes the average power of in 1Hz (aka spot noise).

Then denotes average noise power in bandwidth.

For a 50Ohm resistor at room temperature, noise power in a 1Hz bandwidth is

Noise voltage is

has no physical meaning.

### Effect of transfer function on noise

Noise is filtered just like any other signal. can be a desired signal, or noise, or interference. If we apply to an LTI system with transfer function , then the output spectrum is

### Device noise

We model the noise of electronic devices by voltage and current sources. This allows us to analyze circuit noise using standard circuit analysis techniques.

#### Thermal noise of resistors

Ambient thermal energy leads to random agitation of charge carriers in resistors and hence noise.

Noise is modeled as either a voltage source in series with the resistor or current source in parallel with the resistor.

1. Thevenin equivalent: voltage source in series with the resistor with PSD of
2. Norton equivalent: current source in parallel with the resistor with PSD of

If a resistor converts ambient heat to a noise voltage or current, can we extract energy from the resistor? In particular, let’s analyze this circuit.

A diagram of a circuit

AI-generated content may be incorrect.

If and are at the same temperature, no net energy is transferred between them because the power that delivers to is the same as the power that delivers to .

What if is at 0 K? Then draws thermal energy from its environment, converting it to noise and delivering it to .

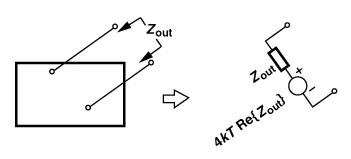
The power delivered to is

Delivered power reaches a maximum when :

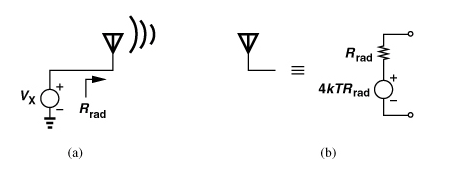
**, called the available noise power, is independent of resistor value and has dimensions of power per Hz. At room temp (), .**

A circuit does not need to contain an explicit resistance of to exhibit a thermal noise density of – the noise of a resistor can be filtered, suppressed, amplified by the surrounding circuit. If a passive circuit dissipates energy, it must contain a physical resistance (since capacitors and inductors store energy, they do not dissipate) and must therefore produce thermal noise. Loosely, “lossy circuits are noisy.”

**This theorem follows: if the real part of the impedance seen between two terminals of a passive (reciprocal) network is equal to , then the PSD of the thermal noise seen between the terminals is .**

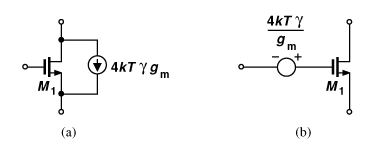
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This isn’t limited to lumped circuits. A transmitting antenna dissipates energy by radiation according to , where is the “radiation resistance. As a receiving element, the antenna generates a thermal noise PSD of .



#### Noise in MOSFETs

The thermal noise of MOS transistors operating in the saturation region is modeled as either a current source tied between source and drain or a voltage source in series with the gate.



Gate, drain, and source exhibit physical resistances that, in a good design, are much less than the channel noise.

MOS devices suffer from “flicker” or “1/f” noise. Again, this can be either a voltage source in series with the gate or a current source in parallel with the channel.

is a process-dependent constant. is typically lower for PMOS because charge in the channel tend to travel well below the silicon-oxide interface and thus suffer less from “surface states” (dangling bonds).

**1/f means that slow noise components have larger amplitude. The choice of the lowest frequency in the noise integration depends on the time scale of interest and/or the spectrum of the desired signal.**

1/f noise PSD intercepts thermal noise PSD at a frequency called the 1/f noise corner frequency (call it ).

We want to be as low as possible.

**1/f noise impacts RF because nonlinearity or time variance in mixers and oscillators translate the 1/f spectrum to RF range.**

### Representation of noise in circuits

#### Input-referred noise

In the lab, we can measure noise only at the output of a circuit. However, it is not fair to compare output noise between different circuits since higher gain means higher noise; that is why we want to refer noise to the input.

Input-referred noise is modeled by a series voltage source and a parallel current source. These two noise sources are necessary and sufficient but often correlated.

A diagram of a circuit

AI-generated content may be incorrect.

Model A is the noisy circuit, and model B is the noiseless equivalent with all noise referred to the input.

To calculate , short the input port of the noisy circuit, calculate the output noise voltage, and divide by voltage gain.

To calculate , open the input port of the noisy circuit, calculate the output noise voltage, and divide by transimpedance gain.

Computing input-referred noise sources is challenging at high frequencies. That’s why we typically use NF.

#### Noise figure (NF)

How does SNR (signal power to noise power) degrade as the signal travels through a circuit? If a circuit contains no noise, then output SNR is equal to input SNR. To quantify the noise of a circuit, we define

How do we calculate ? ? To illustrate, let’s calculate the NF of a LNA.

A diagram of a radio signal

AI-generated content may be incorrect.

The LNA is driven by a source with impedance . This source has noise power . The LNA is represented by a noiseless circuit with input impedance , voltage gain , and output noise .

Both signal and source noise experience an attenuation of when appearing at the LNA input. Therefore

**This gives us a lot of intuition. NF of the LNA is equal to**

1. **Total output noise normalized by output noise from the source**
2. **1 + input-referred output noise normalized by source noise**

**Note: NF depends on source impedance. NF must be specified w.r.t. the source impedance (typically 50Ohm).**

As in this case, NF is typically specified for a 1Hz bandwidth at a given frequency and called “spot noise figure”.

These derivations are based on voltage squared quantities, which means they still hold even if no actual power is delivered. (For example, if , no power is delivered to the LNA.) This is a critical difference between modern RF design and traditional microwave design.

**Example: what is the NF of w.r.t. a source impedance of ?**

A diagram of a circuit

AI-generated content may be incorrect.

If you zero out , . , so source noise at the output is . Then

NF is minimized by maximizing . If for impedance matching, NF cannot be less than 3dB.

#### NF of cascaded stages

Friis’ equation: if you have stages in a signal processing chain, the overall noise figure is

is the NF of stage w.r.t. the output impedance of the previous stage.

is the available power gain of stage .

Available power gain is equal to available power at the output divided by available power at the input. Available power is defined as power delivered if source and load are matched.

Friis’ equation says that the NF contribution of a stage is divided by the available power gains of the preceding stages. If these are gains, then its contribution is shrunk. If these are attenuations, then its contribution is amplified. In practice, stages generally amplify, which means that the NF of the first stage is most important.

#### NF of lossy circuits

The NF of a lossy circuit is equal to its loss/attenuation.

Example: what is the NF of the cascade of a BPF with loss and an LNA with ?

## Sensitivity

Sensitivity is the minimum signal level a receiver can detect with “acceptable quality”, i.e. sufficient SNR. Sufficient SNR depends on modulation and BER requirement.

To calculate sensitivity, start with the NF of the entire receiver chain:

is the input signal power and is the source resistance noise power (both per unit BW).

Typically, the source is the antenna, and the receiver input impedance is matched to the antenna, so it’s common to express everything in W/Hz or dBm/Hz.

Assuming a flat PSD for signal and noise and a signal bandwidth of ,

This equation expresses the weakest detectable signal for a minimum .

Converting everything to dBm and dB by taking 10log10 of both sides,

Remember, is the source noise power at the DUT input, taking into account the interface between source and DUT. Since the antenna and receiver are typically matched, .

In dB,

The first three terms are the total integrated noise of the system – the noise floor.

## Dynamic range

DR is the ratio of the maximum input level a receiver can “tolerate” and the minimum input level it can detect (sensitivity).

Different applications have different definitions of DR.

In ADCs, DR is full-scale input level divided by input level at which SNR = 1.

In RF design, there are two definitions: dynamic range and spurious-free dynamic range. In both, the low end is sensitivity.

**Dynamic range:** the top end is defined as the maximum desired signal level. For example, when a cell phone is close to the base station, signal strength is very large. The receiver must be able to receive this signal without distortion. DR can be very large since the receiver can detect and attenuate large signal.

* Bottom: limited by noise floor
* Top: limited by compression of the desired signal
* Depends only on the signal and receiver noise

**Spurious-free dynamic range:** the top end is limited by IM3 products from two-tone interferer test. Two-tone tested is conducted, and when the power of one IM3 tone is equal to the integrated noise, the power in one interferer tone at the input is the top end. SFDR measures the receiver’s ability to detect a small signal in the presence of interferers. Generally, SFDR is much lower than DR.

* Bottom: limited by noise floor
* Top: limited by IM3
* Depends on signal, receiver noise, and receiver IIP3

A diagram of performance and sensitive

AI-generated content may be incorrect.

SFDR calculation.

@ device input, the IM3 equation is

At max level, IM3 is equal to integrated noise.

SFDR is minus sensitivity.

**Remember, we saw IM3 can corrupt a signal even if gain is not significantly compressed. Similarly, SFDR is much more limited by IM3 in the presence of two interferers than by gain compression (P1dB) in the presence of one interferer.**

## Passive impedance transformation

At RF, we often use passive networks (matching networks) to transform impedances: low impedance to high, high to low, real to complex, complex to real. These networks are usually standalone, not integrated.

### Quality factor (Q)

Q indicates how close to ideal an energy-storing device is. Ideal = Q is infinite.

Nonideal devices dissipate energy due to parasitic resistances. These can be modeled either in series or in parallel with the device.

Capacitors:

Inductors:

### Series-to-parallel conversion

Consider these two circuits. Which values make them equivalent?

A black and white diagram of a molecule

AI-generated content may be incorrect.

Equate impedances.

Substitute .

Then

You can rearrange the first equation to show that .

These two circuits are not equivalent over all frequencies, e.g. at low frequencies, the series circuit approaches an open while the parallel circuit approaches . But it can be true for a narrow frequency range.

Substituting the two previous equations,

When , true for a finite frequency range,

Series-to-parallel conversion retains the value of while boosting by .

Conversely, parallel-to-series conversion reduces by .

The same is true for RL sections.

### Basic matching networks

## Scattering parameters

# Communication concepts

## Analog modulation

Modulation converts a baseband signal to a passband signal. It varies certain parameters (amplitude, phase, frequency) of a sinusoidal carrier according to the baseband signal.

**Amplitude modulation:**

In AM, the amplitude of the carrier is modulated by , the baseband signal.

is called the “modulation index” and has units of .

is real, so is symmetric around .

**Phase modulation:**

In PM, the excess phase of the carrier is linearly proportional to .

**Frequency modulation:**

Remember, frequency is the derivative of phase w.r.t. time: .

Phase is the integral of frequency: .

In FM, the excess frequency of the carrier is linearly proportional to .

**Narrowband FM approximation:**

This is a special case of FM that is useful in the analysis of RF systems.

Let .

The BW of the AM signal is equal to .

In general, the BW of the PM and FM signals will be much larger because the dependence on is nonlinear.

However, if we assume – narrowband FM approximation – we will see that BW is still approx. .

Use small signal approximation and the trig identity

Use another trig identity

From these results, you might be tempted to say that AM sidebands have the same sign while FM sidebands have opposite signs, but this isn’t true. What’s true is that AM sidebands always add up in phase with the carrier – and thus modulate the amplitude – while FM sidebands always add up to be perpendicular to the carrier – and thus mostly modulate phase.

Using phasor analysis (which is just working with baseband equivalents), we can write

The FM component is imaginary; it is always orthogonal to the carrier.

The AM component is real; it is always in the same or opposite direction of the carrier.

## Digital modulation

### ASK, PSK, FSK

Both of these signals are modulated by multiplying with a random bit sequence (sequence of rectangular pulses).

They have the same spectrum (sinc^2) except ASK has impulse at DC, .

### Intersymbol interference

Distortion that arises because of filtering (LTI systems).

A signal cannot be both time-limited and band-limited. When a single rectangular pulse – of infinite bandwidth – passes through a LPF, the output exhibits an exponential tail that grows as filter BW decreases. Because the output is band-limited, it must extend to infinity in time.

When a random sequence of ones and zeros passes through a LPF, each bit is corrupted by the tail of the previous bit(s). This leads to a higher error rate.

Tradeoff: reduce BW to reduce integrated noise, but increase ISI.

Any system that removes part of the spectrum of a signal introduces ISI.

A general random binary sequence can be modeled as

is one pulse of width . .

Nulls at , main lobe width of , side lobes of width .

**Pulse shaping**

To avoid ISI, shape the baseband pulse to have less BW. Sharp transitions = large BW.

The ideal pulse is the sinc, which yields a brickwall spectrum with bandwidth .

If you sample every , there is zero ISI because each pulse has nulls at the peak of every other pulse. This is called Nyquist signaling.

In practice, sinc pulses are difficult to generate so we use approximations.

This pulse is created from a “raised-cosine” filter, so-called because the spectrum looks like a raised cosine.

is the roll-off factor. , sinc. , much wider spectrum. Typically

### Signal constellations and EVM

ASK and PSK signals have one basis function. FSK has two.

Basis functions must be orthogonal – equivalently, zero correlation or inner product is 0.

Constellations are useful for comparing detectability of different modulation schemes, and they provide a quantitative measure of signal quality: EVM is equal to the average power of the error vectors (error between ideal constellation point and measured point) normalized to the signal power.

EVM is typically normalized to percentage. Alternatively, you can convert to SNR.

### Quadrature modulation

In quadrature modulation, we choose the orthogonal basis functions and (called the “quadrature phases” of the carrier).

The inner product of two continuous complex-valued functions on the interval is defined as

They are orthogonal if .

Another way to look at it is from the baseband equivalents (or phasor analysis).

In quadrature PSK (QPSK), we subdivide the binary data stream into pairs of two consecutive bits and impress those bits onto the quadrature carriers:

The baseband equivalent is (let )

Which we can represent as . This means

Converting back to RF equivalent,

That is,

And

Since and are orthogonal, the bits can be separated at the receiver without corruption.

and are called the I and Q (for in-phase and quadrature) baseband signals, and is one symbol.

The original bitstream is composed of pulses of duration , so for the same bitrate, the symbol streams are composed of pulses of duration (symbol rate is ½ of bitrate). This halves the occupied BW (or equivalently, doubles bandwidth efficiency). The original bitstream (with pulse-shaping) has BW . The symbol streams, which overlap in spectrum, have BW .

The main drawback of QPSK (and other quadrature modulation schemes) is that with baseband pulse shaping, the (variable-envelope) signal has very high PAPR which requires linear (and inefficient) power amplifiers.

### GMSK and GFSK

FSK modulation schemes 🡪 constant envelope 🡪 PA can be nonlinear 🡪 high PA efficiency.

Can be implemented by applying baseband pulses to a VCO (frequency is tuned by voltage). Square baseband pulses result in wide spectrum.

A common method of shaping the pulses is called Gaussian filtering: apply a filter to the pulses whose impulse response is a Gaussian pulse.

Gaussian minimum shift keying (used in GSM):

Gaussian frequency shift keying (used in Bluetooth):

is the impulse response of the Gaussian filter.

To implement GMSK/GFSK in IQ transmitter,

### QAM

16QAM = 4 bits/symbol = 4x BW efficiency

64QAM = 6 bits/symbol = 6x BW efficiency

Denser constellation, worse detectability

Higher PAPR

### OFDM

Multipath propagation: the propagation of EM waves from Tx to Rx through multiple paths. The waves arrive at Rx with different delays – “delay spread”. Even if this doesn’t result in destructive interference, they may lead to ISI because you receive multiple copies of the same signal that are shifted in time.

Another perspective from the frequency domain: multipath introduces memory, which filters the signal. Anything that removes part of the signal spectrum causes ISI.

ISI is worse as delay spread increases or as bit rate increases, because symbol duration becomes shorter.

OFDM mitigates the effect of delay spread by transmitting symbols in parallel on multiple subcarriers. If a single-carrier signal has a symbol rate of , then an OFDM signal with subcarriers splits the symbol stream into substreams of rate . Each subcarrier has the original BW.

Total BW and date rate remain the same, but OFDM (multicarrier) is much more resistant to multipath effects because it is composed of low-rate data streams.

Frequency domain perspective: multipath still occurs, BUT the impact on each individual subcarrier is much less.

The downside of OFDM is that adding orthogonal subchannels results in a waveform with very high PAPR even if the modulation scheme of each subchannel has low PAPR.

For OFDM, PAPR is approx. equal to if is large.

## Spectral regrowth

Only variable envelope signals experience spectral regrowth (ACLR, ACPR) from odd-order nonlinearity.

Constant envelope (but modulated) signal:

The second term represents gain compression. The first term is a modulated harmonic, but it’s very far from the desired signal.

Variable envelope signal:

The spectral regrowth term is , which is typically 3 times wider in bandwidth than the desired signal. This spills into adjacent channels.

## Mobile RF communications

## Multiple access techniques

### Time and frequency division duplexing

TDD: same frequency band for Tx and Rx; the system transmits and receives at different times.

* One BPF and one RF switch to alternate between Tx and Rx = low loss
* Tx does not interfere with Rx
* Strong signals by nearby mobiles will desensitize receivers

FDD: two different frequency bands for Tx and Rx, using filters to isolate the two paths and allowing simultaneous transmission and reception. Two transceivers cannot communicate directly. In wireless, the frequency translation is performed by the base station.

* Isolates receivers from transmissions by nearby mobiles
* Requires a duplexer (two BPF). Duplexer must suppress strong Tx signal from leaking into Rx path, but finite suppression means self-desense. Furthermore, due to tradeoff between quality factor and loss, duplexer loss tends to be high.
* Tx transients (PA on/off, LO transient) leak to adjacent channels (in TDD, transients can be timed to end before switching to Tx).

### Frequency- and time-division multiple access

FDMA:

* The frequency band is partitioned into many channels, each of which is assigned to one user. When paired with FDD, each user is assigned two channels (one Tx, one Rx).
* Relatively simple, so it was used in early cellular networks called “analog FM” systems
* On its own, does not grant much capacity (maximum number of simultaneous users)

TDMA:

* The same band is available to each user but at different times. Each transceiver is periodically given access for a time slot of duration . The overall period consisting of all slots is called a frame (). Every seconds, each user can access the channel for seconds.
* This means the device must buffer data while it is waiting for its turn and then transmit as a burst during its one time slot (TDMA burst). This requires digitizing the data, which also allows for compression and coding.
* PA only needs to be on for part of the time
* Compressing data = more bandwidth and time-efficiency and higher capacity
* Even with FDD, TDMA bursts can be timed s.t. Tx and Rx are never simultaneously enabled
* Requires A/D conversion, digital modulation, time slot and frame synchronization
* Most TDMA systems also incorporate FDMA – multiple channels, time-shared among users

### Code-division multiple access

## Wireless standards

Common specifications:

1. Frequency bands and channelization
2. Data rate(s) – some standards specify a constant date rate, while others allow for lower data rates in adverse channel conditions
3. Antenna duplexing method, e.g. FDD, TDD
4. Modulation – sometimes, different modulation schemes are used for different data rates
5. Tx output power – some standards require variable Tx output power to save battery and to avoid near/far effects
6. Tx EVM and spectral mask – also, spurs and harmonics
7. Rx sensitivity – usually in terms of a maximum bit error rate. Usually, higher (better) sensitivity for lower data rates.
8. Rx input level range
9. Rx tolerance to blockers – defines the performance of the receiver in the presence of large interferer(s) with a small desired signal (e.g. 3dB above sensitivity). Can be one interferer (blocking requirement) or two interferers (intermodulation requirement). When the interferers are at the specified power, BER must be below maximum bit error rate.

### GSM

* TDMA + FDD system – Tx and Rx are on different channels, and each channel is time-multiplexed among 8 users.
* 200kHz channel
* GMSK modulation
* 271 kb/s per user
* Tx and Rx time slots are offset by ~1.73ms so they don’t operate simultaneously

A 25MHz frequency band can support 1000 users, i.e. the system’s capacity is 1000 users.

**Blocking requirements:** Desired signal at reference sensitivity+3dB with unmodulated tones applied one-at-a-time in discrete increments of 200kHz from the desired channel.

**A diagram of a band blocker

AI-generated content may be incorrect.**

**Intermodulation requirements:** Desired signal at reference sensitivity+3dB; a tone applied at 800kHz offset and a modulated signal at 1.6MHz offset. Each interferer is at -49dBm.

A diagram of a graph

AI-generated content may be incorrect.

**Adjacent-channel interference:**

**GSM Rx sensitivity is -102dBm. To detect GMSK with acceptable BER of , we need SNR of 9dB. What is the required NF?**

**To satisfy blocking, what is the required P1dB?**

Since the out-of-band blockers are attenuated by the front-end filter, we expect the largest in-band blocker, -23dBm @ 3MHz offset, to dominate this requirement. To ensure very little gain compression, P1dB needs to be roughly -15dBm.

While the 3MHz offset blocker dominates P1dB, smaller offset blockers are specified because of reciprocal mixing.

The blocking requirements are very difficult to satisfy, so GSM specifies “spurious response exceptions”, allowing 6 in-band frequencies and 24 out-of-band frequencies to be relaxed to -43dBm. However, this doesn’t ease the compression and phase noise requirements since the number of blocking frequencies >> 6.

**Estimate the required IIP3 to meet the intermodulation requirement.**

Signal is at -99dBm. Noise floor is

Required SNR is 9dBm, so the sum of noise and IM3 must be equal to -108dBm, which means IM3 must be at -111dBm.

This corresponds to a P1dB of -28dBm. The P1dB requirement for blocking is -15dBm, so Rx linearity is driven by single-tone blocking rather than IM.

#### EDGE

### IS-95 CDMA

* Qualcomm
* Direct-sequence CDMA + FDD
* Uplink: 9.6 kb/s baseband data 🡪 spread to 1.23MHz 🡪 OQPSK modulation
* Downlink: QPSK
* Mobile must use a power-efficient modulation, while the base station transmits many channels simultaneously and thus needs to employ a linear PA regardless of modulation
* In both UL and DL, IS-95 requires coherent detection – transmit a strong “pilot tone” (unmodulated carrier) at the beginning of communication to establish phase synchronization

**Power control:**

CDMA requires power levels received at the base station to differ by no more than ~1dB from all the different mobiles. Power control starts with open loop and settles into closed loop.

Open loop: mobile measures the DL signal power and adjusts UL power so the sum is -73dBm. Assuming UL and DL channels have the same loss, power at the base station is well-controlled.

Closed loop: in reality, UL and DL channels are at different frequencies, so the two paths may experience different fading. The base station measures the power level from the mobile and sends a feedback signal requesting power adjustment. This command is transmitted every 1.25ms to ensure timely adjustment in the presence of rapid fading.

**Frequency diversity:**

Spreading spreads the original spectrum to a wider BW. This wider BW is more unlikely to experience total suppression due to multipath fading. It will experience frequency-selecting fading, but the effect of this is mitigated after dispreading.

**Time diversity:**

### WCDMA

# Transceiver architectures

## General considerations

The most important constraint in wireless communications is the limited channel BW for each user. Shannon’s theorem says the theoretical capacity (maximum achievable data rate) in a channel with bandwidth and AWGN is

This dictates the use of sophisticated baseband processing techniques like coding, compression, and BW-efficient modulation.

The narrow channel BW also affects transceiver design: transmitter must avoid leaking power to adjacent channels, and receiver must be able to detect the narrowband signal while rejecting in-band and out-of-band interferers.

In particular, the linearity of the Rx must be good enough to accommodate interferers without experiencing significant compression or intermodulation.

**Is it possible to filter interferers and thus relax the linearity requirement of the Rx?**

Let’s use an example. A 900-MHz GSM Rx must tolerate an alternate adjacent blocker 20dB higher than the desired signal. Calculate the Q of a second-order LC filter required to suppress the blocker by 35dB.

A second-order RLC tank (parallel components) has impedance

The magnitude response squared is

Assume the resonance frequency is .

At the blocker frequency – 400kHz offset from 900MHz, or 900.4MHz – we want

**Channel selection and band selection:**

The filter in the previous example is an example of channel selection – it selects the desired signal channel and rejects interferers in other channels.

* All of the Rx stages that precede channel selection must be sufficiently linear to avoid compression and intermodulation
* Channel selection at the RF channel frequency requires prohibitively high Q and thus must be deferred to a later stage. BPF Q may be approximated as center frequency divided by 3dB BW.

Typically, there is a band-select filter at the RF channel frequency that rejects interferers outside of the receive band. That is why we distinguish between in-band and out-of-band interferers.

The dual of this issue on the Tx side is that there is no channel selection to filter spectral regrowth, which means modulation and PA linearity must be designed to minimize leakage.

There is a tradeoff between filter loss and selectivity: higher selectivity = higher order = more cascaded sections = higher loss. Typically, the loss of the band-select filter is prioritized, which means that Rx linearity is important.

**Tx-Rx feedthrough**

In full-duplex standards, Tx and Rx operate concurrently (like CDMA, which require continual power control). There are two leakage concerns:

* Tx signal leaking to Rx – for a typical 50dB attenuation, the Tx signal at the Rx port is -20dBm
* Noise generated by the Tx, including nonlinear products, at Rx frequencies leaking from Tx port to Rx port

Typical attenuation from Tx port to Rx port is 50dB for both leakage mechanisms.

## Receiver architectures

### Basic heterodyne Rx

### Modern heterodyne Rx

### Direct-conversion Rx

Advantages over heterodyne:

* No image
* Channel selection is performed by LPF, which can be realized on-chip as active circuit topologies with relatively sharp cut off
* Fewer mixing spurs
* Easily lends itself to integration – only the BPF needs to be off-chip. This means the LNA/mixer interface can be optimized without requiring 50Ohm impedance.

#### LO leakage

The LO the drives the mixer couples to the Rx antenna (through parasitic device capacitances and through the substrate) and may desensitize other receivers.

This can be mitigated if the LO has differential outputs that couple identically to the antenna (symmetric layout).

This occurs in heterodyne Rx, too, but because the LO frequency is outside of the band, it is attenuated by the emitting Rx.

#### DC offsets

LO self-mixing (LO leaks to antenna and is amplified by the Rx chain being downconverted) results in DC offsets. This must be handled carefully as a large DC offset may saturate baseband circuits. This problem also occurs in heterodyne Rx having zero second IF.

I and Q DC offsets are generally unequal. Let’s say the summation of the leakage components from the quadrature phases of the LO is equal to leakage components from the quadrature phases of the LO is equal to

After downconversion in the mixer, the DC components are

Where and are the gain and phase shifts through the LNA and mixers.

Typically, DC offset is cancelled via a programmable current source, tied to the mixer output node (mixer outputs a current), that sinks a current that drives DC offset down to zero.

This programmable current source is calibrated across circuit configurations. The calibration settings are loaded per configuration. For example, LO leakage can change due to

* Different LO frequency = different coupling strength
* Different LNA gain states = different reverse isolations

#### Even-order distortion

Any Rx with final zero IF is sensitive to even-order distortion.

**Beat component:**

Typically, circuits are differential – they have odd symmetry specifically to suppress even-order distortion. However, imperfections in the symmetry result in finite even-order terms. The mixer usually dominates this nonideality.

The low-frequency “beat” is generated by :

has units of .

Like IM3, IM2 is quantified via IIP2.

Everything in dBm:

**Variable-envelope signal or blocker:**

When a signal or blocker has amplitude modulation, second-order nonlinearity results in a modulated baseband term that corrupts the downconverted signal.

The component of interest is . What is the corruption due to the signal itself?

The downconverted signal can be expressed as . Then SNR is

This problem is worsened if there is a large AM blocker, since SNR becomes

is the envelope of the blocker.

#### Flicker noise

Gain in the LNA/mixer chain is typically limited to 30dB because of linearity requirements. This means the downconverted is still quite small and thus susceptible to noise – especially flicker noise – in the baseband circuits.

In the downconverted spectrum,

* is half of the RF channel BW
* is the thermal noise at the end of the baseband chain
* is the flicker noise

A graph of a function

AI-generated content may be incorrect.

We want to calculate the flicker noise penalty – how much overall noise power increases from just alone. This requires integrating the flicker noise spectrum. The upper limit of integration is . You can’t choose 0 as the lower limit – it’s undefined. The lower limit must be chosen s.t. frequencies below the lower limit effectively have no impact within the timescale of interest (e.g. the burst). The effect of frequency components below the lower limit is effectively a DC offset, which can be handled deterministically.

Let be the total noise power; let be noise power from alone.

Flicker noise penalty is a function only of and the integration limits.

In a good design, the total noise is dominated by the antenna, LNA, and mixer – that is, we want the gain through these stages to be high, raising , and lowering .

**Examples:**

An 802.11g receiver has .

A GSM receiver has .

The data rate for GSM is 271 kb/s. Let’s assume the lower limit of integration is equal to 271 kb/s divided by 1000.

The equation for flicker noise penalty was derived assuming , which is not true in this case.

A diagram of a function

AI-generated content may be incorrect.

Standards with narrow channel bandwidths are more susceptible to flicker noise and may necessitate the use of a low-IF architecture instead of zero-IF.

#### IQ mismatch

Quadrature modulation requires the basis functions to be orthogonal. The I and Q signal paths must be perfectly balanced, and the quadrature phases of the LO must be offset by exactly 90 degrees. I/Q asymmetry (mixer, baseband) and errors in the phase shift circuit break orthogonality, causing signal corruption.

IQ mismatch tend to be larger in direct conversion compared to heterodyne because of the higher LO frequency:

* For a given delay mismatch, phase error is directly proportional to frequency
* Higher operating frequency requires smaller device dimensions, leading directly to larger transistor asymmetry

Because of the phase mismatch, I and Q are no longer orthogonal, and I and Q experience “crosstalk” – I corrupts Q and vice versa.

It’s straightforward to measure and compensate IQ mismatch. Mismatches can be measured via an RF tone.

Amplitude and phase can be calculated via SPDFT.

#### Mixing spurs

Interferers at harmonics of the LO will mix with the harmonics and corrupt the channel.

### Image-reject Rx

### Low-IF Rx

## Transmitter architectures

Modulation, upconversion, and power amplification.

Most systems use quadrature baseband signals and IQ upconversion.

One way to implement pulse shaping is to run the baseband data through an address generator, which produces a digitized pulse. This is then converted to analog.

A diagram of a function

AI-generated content may be incorrect.

### Direct-conversion Tx

* Directly translates baseband signal to channel frequency. Compact solution and relatively “clean” output compared to other architectures.
* PA must deliver a lot of power – it has large transistors to carry high currents and thus exhibits a large input capacitance. Typically, a driver amplifier is interposed b/w the upconverter and the PA to serve as a buffer.
* Matching network so ensure maximum power delivery

A diagram of a network

AI-generated content may be incorrect.

#### IQ mismatch

In Rx, Razavi models the IQ mismatch as balanced – i.e. balanced across I and Q paths. In Tx, Razavi models the IQ mismatch as unbalanced – all on the I path. I think the end result should be the same except for a phase shift.

Upconverted:

Ideal downconversion:

Again, phase mismatch results in cross-talk.

What is the EVM in the presence of IQ mismatch?

An alternative way to quantify IQ mismatch is through the unwanted sideband (aka residual sideband aka image).

Using phasor analysis,

Then the ratio of desired signal power to RSB power is

The total power of signal + RSB is independent of phase mismatch.

**How do we estimate and ?**

Razavi proposes a method using only power measurements.

First, estimate . Apply to I path with zero input to Q path, then apply to Q path with zero input to I path.

After compensating , estimate . Apply to both I and Q. Then we have the plus-minus terms

Using phasor analysis,

Let .

From the previous section, when we apply a complex sinusoid, we have

Then

DC offsets in the baseband may affect the accuracy of IQ mismatch calibration. Carrier leakage may also require the removal of DC offsets prior to IQ mismatch calibration.

#### Carrier leakage

Because of DC offsets in the baseband circuitry and in the baseband ports of the upconversion mixers, the output signal contains a fraction of the unmodulated carrier.

Then

Undesirable consequences:

* Impacts EVM by shifting the signal constellation
* May impact power control if the digital baseband signal is backed off (e.g. for power control to avoid near-far effect)

Razavi proposes a feedback loop, using a least-mean square algorithm (stochastic gradient descent) or exhaustive search to find the DC offset cancellation settings that yield the lowest leakage.

A diagram of a circuit

AI-generated content may be incorrect.

#### Mixer (baseband) linearity

If the baseband signal experiences nonlinearity of ,

Note: the term is identical to spectral regrowth caused by RF nonlinearity.

The new term caused by baseband nonlinearity is

In most cases, PA linearity dominates.

#### Tx linearity

#### Oscillator pulling

# Extra notes

Math is nearly identical to direct-conversion Rx.

Upconverted signal:

Ideal downconversion:

Again, phase error causes cross-talk.

What is the EVM?

Error is :

Then EVM is .

An alternative way to quantify IQ mismatch is through RSB/image. Let . Then

Separate by frequency:

Use phasor analysis.

For the unwanted sideband,

For the desired signal,

Use phasor analysis – translate sin to cosine and add them up using vector addition.