EE40458
Nonlinearity & Noise

Yet another lecture from the road
Recap & Perspective

• An overview – where have we been recently:
  – General linear system theory
    • S-parameters, but also equivalent small-signal models like the hybrid-π model, etc.
    • Give linear response of circuit/system; can use superposition & Fourier analysis to determine output for arbitrary input signal
    • “Gain, phase” at each frequency; no new frequencies, no signal components that were not present in the input signal
    • Strictly applies only for systems governed by linear differential equations (any order, but constant coefficients)
    • Approximately applies to most systems if signals are small (equivalent to approximating function with first two terms (constant plus linear) in Taylor series)
Recap & Perspective (cont.)

• Overview (continued):
  – Nonlinear effects
    • Many important systems are not well approximated with a linear description
    • Examples: almost any component under high power conditions (e.g. heating); amplifiers driven with large inputs (or designed for switching-mode operation for high efficiency); devices for harmonic generation, mixing, or detection (nonlinear response is desired)
    • Our approach: Taylor series expansion of transfer characteristic, resulting in polynomial representation of response
      – We simplified to neglect memory, history-dependent effects (e.g. heating); assumed output depends only on instantaneous input value. More advanced approaches exist to handle this
    • Conclusions: saw harmonic generation (plus DC shift), intermodulation products (e.g. sum & difference frequencies)
    • Figures of merit: $P_{1\text{dB}}$ (gain compression), $P_{\text{IP3}}$ (third order intermodulation)
Nonlinear Figures of Merit - Review

• Gain compression – $P_{1dB}$:
  – Single input tone
  – Map output power (at input frequency) vs. power of input signal
  – Linear theory: output power proportional to input power; nonlinear effects tend to cause saturation:
    \[ v'_o(t) = k_1A \left[ 1 - \frac{3}{4} \left| \frac{k_3}{k_1} \right| A^2 \right] \cos(\omega t) \]
  – Note: input is power in single tone; output power “counted” is only the power at this frequency (i.e., the harmonic power is not included)
Nonlinear Figures of Merit - Review

• Gain compression – \( P_{1\text{dB}} \):
  
  – Definition: Input 1 dB compression point, \( P_{1\text{dB}} \), is input power at which output is 1 dB below the linear case.
  
  
  $$v'_o(t) = k_1 A \left[ 1 - \frac{3}{4} \left| \frac{k_3}{k_1} \right| A^2 \right] \cos(\omega t)$$

  – Note: some data sheets will report the output 1 dB compression point (e.g. the y-axis, rather than x-axis). Depends on intended application.
Nonlinear Figures of Merit - Review

- **Intermodulation –** $P_{IP3}$:
  - Saw complicated relationship with two input tones
  - Figure of merit: for comparison of components – $P_{IP3}$
  - Two input tones, equal amplitude (for figure of merit); small enough that gain compression can be neglected: $v_{in}(t) = A [\cos(\omega_1 t) + \cos(\omega_2 t)]$
  - Output power at close-in intermodulation product frequencies vs. power of input signal. Near $\omega_1$ and $\omega_2$, have:

$$v'_o(t) = k_1A \left[ \cos(\omega_1 t) + \cos(\omega_2 t) \right] + k_3 \frac{3}{4} A^3 \left[ \cos(2\omega_2 - \omega_1) t + \cos(2\omega_1 - \omega_2) t \right]$$

![Diagram of input and output frequencies with in-band and close-in products indicated.](image-url)
Nonlinear Figures of Merit - Review

• Intermodulation – $P_{IP3}$ (continued):
  – Purpose of $P_{IP3}$ figure of merit: quantify relationship between $P_{in}$, $P_{out}$ (at signal frequency), and intermodulation products
  – Map power in “desired” frequencies (first term) to power in intermodulation products (second term)

\[
v_{in}(t) = A\left[\cos(\omega_1 t) + \cos(\omega_2 t)\right]
\]

\[
v_{o}'(t) = k_1 A\left[\cos(\omega_1 t) + \cos(\omega_2 t)\right] + k_3 \frac{3}{4} A^3 \left[\cos(2\omega_2 - \omega_1)t + \cos(2\omega_1 - \omega_2)t\right]
\]

\[
P_{in} \propto \frac{1}{2} A^2
\]

\[
P_{d} \propto \frac{1}{2} k_1^2 A^2 \propto P_{in}
\]

\[
P_{im} \propto \frac{1}{2} k_3^2 \frac{9}{16} A^6 = k_3^2 \frac{9}{4} A P_{in}^3
\]

$P_{in}$ = power at each input tone
$P_{d}$ = power at each desired output tone
$P_{im}$ = power at each intermod output tone
Nonlinear Figures of Merit - Review

• Intermodulation – graphically:
  
  – Graph $P_d$, $P_{im}$ vs. $P_{in}$, usually on log-log scale (all powers in dBm)
  
  – Reminder: dBm = 10*\log10(P/1 mW)

\[
\begin{align*}
  P_{in} &\propto \frac{1}{2} A^2 \\
  P_d &\propto \frac{1}{2} k_1^2 A^2 \propto P_{in} \\
  P_{im} &\propto \frac{1}{2} k_3^2 \frac{9}{16} A^6 = k_3^3 \frac{9}{4} A P_{in}^3
\end{align*}
\]

– $P_{IP3}$: intercept between linear ($P_d$) and intermod ($P_{im}$) terms

– Intercept is “fictitious”; in practice, based on low power data (avoid gain compression); real lab data includes everything...

\[
\begin{align*}
P_d (dBm) &= Gain(dB) + P_{in} (dBm) \\
P_{im} (dBm) &= offset + 3 P_{in} (dBm)
\end{align*}
\]
Intermodulation Analysis

• Measurement:
  – Measure $P_d$, $P_{im}$ at several (low) levels of $P_{in}$
    • Can easily separate $P_d$, $P_{im}$ (on spectrum analyzer) because at different frequencies
    • Use slope of 1 for $P_d$ vs. $P_{in}$ (in dBm)
    • Use slope of 3 for $P_{im}$ vs. $P_{in}$ (in dBm)
    • Find intercept point; input IP3 ($P_{ip3}$, IIP3) or output IP3 (OIP3) can be projected

• Analysis:
  – “Intermodulation Ratio”: $IMR = \frac{P_{im}}{P_{d}} = \left( \frac{P_{in}}{P_{IP3}} \right)^2$
  – Convenient relation:
    • Relate expected intermodulation products from (known) IIP3 and $P_{in}$
    • Find IIP3 given measured $P_{im}$, $P_d$
Noise

• Linear and nonlinear analysis relates output to input stimulus
  – Linear: small signals; non-linear: large signals
• Circuits, systems also produce outputs independent of input: noise
  – Ultimately, noise limits our ability to resolve/recover/process very small signals
• Noise is fundamental – cannot be eliminated; but can be managed
• Sources of noise:
  – Thermal noise: random motion of carriers (electrons, holes) in resistive material
  – Shot noise: cause by random timing of events
    • Current is made of up of flow of electrons, but they have some “jitter” in when they arrive; this generates shot noise
  – Flicker or 1/f noise: trapping/detrapping, often defect or surface related; has ~1/f noise power spectral density
Thermal Noise

• Let’s look at thermal noise in a resistor:

\[ e_n(t) \]

\[ \langle e_n \rangle = 0 \]

\[ \langle e_n^2 \rangle = 4kTBR \]

– Reminder: \( k = 1.38 \times 10^{-23} \text{ J/K} \) (Boltzmann’s const.), \( T \) (in Kelvin)
– Expect average voltage = 0 (no net flow, equal opposite flows)
– But variance ≠ 0; voltage variance proportional to noise power
– B: bandwidth of measurement—how much noise power you see depends on how wide of a frequency range you look at

• Potential problem: noise power goes up as bandwidth increases—no limit?
  – Not really... this conclusion comes from simplistic assumption on carrier statistics. But if \( T \sim 300 \text{ K} \), is good approx. for frequencies to \( \sim 1 \text{ THz} \) or so
Modeling Noise

• For analysis, need equivalent circuit for analysis
• One option: model noisy resistor as noiseless resistor with associated noise source (Thevenin or Norton options):
  \[ \langle e_n \rangle = 0, \langle i_n \rangle = 0 \]
  \[ \langle e_n^2 \rangle = 4kTB, \langle i_n^2 \rangle = 4kTB/R \]
• Can we get power from these sources? Yes, but… Consider:

  – Power available: \[ \frac{e_n^2}{4R} = kTB \] Independent of \( R \)
  – Power transfer? If both resistors at same temperature, net flow = 0 (equal/opposite flows). If at different temperatures, power from hot to cold (attempts to equilibrate the system). You knew that.
Modeling Noise (cont.)

• How about complex impedances?
  – One can show that:
    \[ Z(f) = R(f) + jX(f) \]
    \[ \langle e_n^2 \rangle = 4kT \int_{-B}^{B} R(f) \, df \]
  – No noise from reactances (no loss or dissipation, no noise)

• Numerical example:
  – Noise voltage across at 1 M\(\Omega\) resistor in bandwidth of 100 MHz (e.g. typical oscilloscope input)
    \[ 4kT = 1.6 \times 10^{-20} \text{ J} \quad (T=290 \text{ K}) \]
    \[ \langle e_n^2 \rangle = 4kTBR = 1.6 \times 10^{-20} \cdot 10^6 \cdot 10^8 = 1.6 \times 10^{-6} \text{ V}^2 \]
    \[ \text{rms voltage} : \sqrt{\langle e_n^2 \rangle} = 1.26 \times 10^{-3} \text{ V} \]
  – Can see why oscilloscopes have minimum 2 mV/div scales...anything smaller is just noise
Excess Noise

• Many components exhibit additional noise, beyond the thermal contributions
  – Often modeled as if it were thermal noise, but with modified parameters (i.e., fudge factors)

• Two common approaches:
  – Circuit analysis approach (typical):
    • Introduce $R_n$ as fudge factor, $\langle e_n^2 \rangle = 4kTBR_n$
    • $R_{\text{noiseless}}$ is actual $R$ value for circuit
    • For pure thermal noise (no excess noise), $R_{\text{noiseless}}=R_n$
  – System analysis approach:
    • Use $T$ as fudge factor: $\langle e_n^2 \rangle = 4kT_n BR$
    • $T_n$ no longer “thermometer” temperature; if excess noise present, $T_n>T$
    • Concept of “noise temperature” is common, and what we’ll (mostly) use
Noise Temperature Example

• An example: antenna
  – At resonant frequency, antenna impedance ($Z_{ant}$) is resistive
  – $Z_{ant} = R_{ohmic} + R_{rad}$
    • $R_{ohmic}$: from loss in the conductors, a “real” resistance
    • $R_{rad}$: accounts for conversion from input power to radiated power (think of antenna as broadcasting)
  – Temperatures?
    • $R_{ohmic}$: at physical temperature of the antenna—electrons bouncing around in conductors due to random thermal motion
    • $R_{rad}$: at an “effective” temperature, $T_A$
    • $T_A$ is “fudge factor” to allow us to make output noise of antenna match the power actually received
    • Fun fact: “Cosmic background radiation” – 1978 Nobel prize – measured $T_A \sim 3K$, when expected to be 0 (dark sky)
Noise in Two-Port Networks

• So far, everything has just been about how much noise a one-port circuit makes. But two-port networks are more useful—have inputs, outputs

• Basic idea: two-port network does some function (amplify, mix, etc), but also adds some noise
  – Schematically:

  – $T_E$: effective noise temperature of two-port network
    • Is a function of $Z_s$, frequency; characterizes the network, not $Z_s$
Noise in Two-Port Networks, cont.

- To include effects of both two-port and termination, use “operating temperature”, $T_{op}$
  - $T_{op} = T_E + T_s$
  - Adding temperatures: same as adding powers. Assumes no correlation between noise sources

Characterization? Two-port network has s-parameters, plus $T_E$
Noise Factor

• Another common way to characterize the noise added by a two-port network is the noise factor and noise figure

• Two equivalent definitions:
  – Definition #1: \( \text{Noise Factor} \equiv \frac{\text{Input SNR}}{\text{Output SNR}} \bigg|_{T_s=T_0=290\,\text{K}} \)

\[
F = \frac{S_{in}/N_{in}}{S_{out}/N_{out}} = \frac{S_{in}/N_{in}}{G_A S_{in}/N_{out}} = \frac{N_{out}}{G_A N_{in}}
\]

\( T_0 = 290\,\text{K} \) (std. temp.)
\( S_{in} \) = available signal power
\( N_{in} \) = available noise power
\( S_{out} \) = available output signal
\( N_{out} \) = available output noise
\( G_A \) = available gain: \( S_{out}/S_{in} \)
\( N_A \) = noise added by 2-port (per sandwich)
Noise Factor & Noise Figure

• Alternative view:
  – Definition #2: \[ F \equiv \frac{\text{Actual available noise power (output)}}{\text{Available noise power if two-port was noiseless}} \]

Result:
\[ F = \frac{N_{out}}{G_A N_{in}} = \frac{N_A + G_A N_{in}}{G_A N_{in}} \]

• Can be framed in terms of temperatures:
\[ F = 1 + \frac{N_A}{G_A N_{in}} = 1 + \frac{N_A / G_A}{N_{in}} = 1 + \frac{kT_E B}{kT_O B} = 1 + \frac{T_E}{T_o} \]
  – So providing F is equivalent to providing T

• Technically, F=noise factor

• Noise Figure is more common; F converted into dB
  – \[ NF = 10 \times \log_{10}(F) \]
Noise Figure

• Caution:
  – From definition #1: \[ \text{Noise Factor} = \frac{\text{Input SNR}}{\text{Output SNR}} \mid T_s = T_o = 290 \text{K} \]
  – Looks like noise figure should be how much (in dB) the SNR degrades because of the noise of the two port
  – This is not strictly true: note that this is true only if \(T_s = T_o\)
  – As we saw, \(T_s\) can be an effective temperature with no obvious connection to “thermometer” temperatures (e.g. if signal came from an antenna, etc)

• So noise figure should be thought of as a “test-based metric”
  – In the lab, can test the SNR with \(T_s = T_o\), and find NF
  – In real systems, \(T_s\) is almost never \(T_o\), so the actual SNR change can be quite different
Noise Figure and LNA Design

• How is this related to our LNA design approach?

• Recall:

\[ F = F_{\text{min}} + \frac{4R_n}{Z_0} \frac{|\Gamma_s - \Gamma_{\text{opt}}|^2}{\left(1 + \Gamma_{\text{opt}} \right)^2 \left(1 - |\Gamma_s|^2 \right)} \]

• This shows explicitly how F depends on \( Z_s \) (\( Z_s \leftrightarrow \Gamma_s \))
  – This is important if you’re doing detailed circuit design (e.g. of an amplifier to meet a specific noise figure target)

• But for system design or analysis, the \( Z_s \) is usually already defined and fixed
  – Block diagram-level interconnections; not re-designing the individual components
  – Example: making a system by interconnecting available “50 Ω” components

  • In this case, \( T_E \) is sufficient (if \( T_E \) is for the \( Z_s \) in question)
Noise Figure and Loss

• A special-case two-port is the matched attenuator
• Passive device—just a resistor network. Typically designed to have input & output impedance matched to $Z_o$, with a specified attenuation (e.g., 3 dB. 6 dB. etc).

• What is it’s noise figure?
  – Since network is just passive, at output appears as $Z_o$ termination at physical temperature; $N_{out} = kTB$
  – But regular two-port equations also apply: $N_{out} = kTB \cdot G_A + G_A \cdot N_{added}$
  – Loss=1/Gain, so: $T_E = T_{att} (L-1)$
  – If $T_{att}=T_o$,
    \[
    F = 1 + \frac{T_E}{T_o} = 1 + \frac{T_{att}}{T_o} (L-1)
    \]
    then $F=L$
System Noise Analysis

• Common situation: want to evaluate noise performance of a system consisting of several building blocks cascaded together.

\[ T_E = T_{E1} + \frac{T_{E2}}{G_{A1}} \]

\[ F = 1 + \frac{T_E}{T_O} = F_1 + \frac{F_2 - 1}{G_{A1}} \]

• Overall noise?
  – Can show:  
  
  – Careful: G’s are available power gains, T_E’s must be for actual impedances presented.
System Noise Example

• Consider two amplifiers in cascade:

\[
\begin{align*}
T_E &: 100k \\
G_1 &: 10
\end{align*}
\]

\[
\begin{align*}
T_E &= 100k + \frac{1000k}{10} \\
&= 200k \\
F &= 1 + \frac{T_E}{T_0} \\
&= 1 + \frac{200k}{200k} \\
&= 1.69 \quad \text{(not dB)}
\end{align*}
\]

\[
NF = 10 \log_{10} (F) = 2.3 \text{ dB}
\]

\[
\begin{align*}
T_E &= 100k + \frac{100k}{10} \\
&= 1010k
\end{align*}
\]

\[
F = 4.48, \quad NF = 6.5 \text{ dB}
\]

• And if we reverse the order of the amplifiers?
  – Gain is the same, but what about noise?

  Note: *usually* want the lowest $T_E$ amplifier in front. But not always—the gains also play a role.
Receiver Sensitivity

- For radio receivers, sensitivity is limited by noise floor
- Define: minimum detectable signal (MDS) for given SNR
  - Often choose 0 dB as the threshold (though other choices are possible depending on the system)
- Example:

  ![Diagram of receiver sensitivity](image)

  - Results:
    - \( NF = 8 \text{ dB} \rightarrow F=6.31 \rightarrow T_E = 1540 \text{ K} \)
    - \( T_{op} = 290 \text{ K} + 1540 \text{ K} = 1830 \text{ K} \)
    - \( N_{out} = k T_{op} B G_A; S_{out} = S_{in} G_A; \quad SNR_{out} = \frac{S_{out}}{N_{out}} = \frac{S_{in}}{kT_{op} B} \)
    - Setting \( SNR_{out} = 1 \rightarrow S_{in} = 5.3 \times 10^{-17} \text{ W} = -132.8 \text{ dBm} \)
    - Effective noise floor of the receiver; for reference, thermal noise at 290 K = -174 dBm/Hz
Noise Figure Measurement

• Basic idea: measure output noise power for two different source temperatures
  - From this, can separate contribution from source and from two-port

• “Y factor” measurement: output powers
  - $N_H = k(T_E + T_H)BG_A$
  - $N_C = k(T_E + T_C)BG_A$
  - $Y = \frac{N_H}{N_C} = \frac{T_E + T_H}{T_E + T_C}$; $T_E = \frac{T_H - YT_C}{Y - 1}$
Noise Figure Measurement

• For “hot” and “cold” input terminations, could use resistors at different temperatures
  – For best measurements, want largest possible difference between hot and cold temperatures (cryogenic resistor, hot resistor)
• Inconvenient in practice, often use “noise diode” instead
  – Diode off (no bias): room temperature resistor (thermal noise)
  – Diode “on” (biased in reverse breakdown): avalanche breakdown process is very noisy, acts like “hot” resistor
    • Characterized by the “excess noise ratio”
    \[
    \text{ENR} = 10 \log_{10} \left( \frac{T_H - 290}{290} \right)
    \]
    • Typical ENR ~15 dB; \( T_H \approx 9461 \text{ K} \) (Solar surface ~6000 K). Much bigger temperature difference than possible using “thermometer” temperatures
  – A caveat: we know that amplifier noise figure depends on \( \Gamma_s \): so if diode has different impedance in “on” and “off” states, measurement can be off—often use “low ENR” diode to avoid this. Just a regular “high ENR” diode, followed by an attenuator
Dynamic Range

• For systems, very important consideration
• Dynamic range: range of input signal amplitudes for which the system has “acceptable” performance
  – System-level considerations dictate what counts as “acceptable” – some systems can tolerate more distortion than others, etc.
  – Limited for large input signals by nonlinearities
  – Limited for small input signals by noise
  – So we’ll be combining the last few topics together
Dynamic Range Definition

• Basic notional system picture:

• With this framework in mind, can define dynamic range:

\[
DR = \frac{\text{max. usable input power}}{\text{min. usable input power}} = \frac{P_{\text{max}}}{P_{\text{min}}}
\]

• Usually expressed in dB—since these are powers, \(10 \log_{10}(DR)\)
Spur Free Dynamic Range

• One common (but not universal!) choice for “acceptable” performance: spur free dynamic range (SFDR)
• SFDR = DR_f (same thing, different terminology)
  – For SFDR: $P_{\text{min}} =$ minimum detectable signal (MDS)
    • But what is the MDS? Smallest $P_{\text{in}}$ that will provide a specified SNR at the output. Often this “reference” SNR=1
  – For SFDR: $P_{\text{max}} =$ input signal level at which 3rd-order in-band products are equal to $P_{\text{min}} (=\text{MDS})$
• Easier to understand (I think) as a picture
Spur Free Dynamic Range

- SFDR in pictures:
  (think of 2-tone intermod measurement)

- $P_{\text{min}} =$ minimum detectable signal (MDS)
- $P_{\text{max}} =$ input signal level at which $3^{rd}$-order in-band products are equal to MDS

- Basic idea: intermodulation is never larger than the noise—no “spurs”
  - In microwave-speak, “spur” is short for “spurious signal”. Not poking horses.
Spur Free Dynamic Range

• Analytically: using \( IMR = \frac{P_{im}}{P_d} = \left( \frac{P_{in}}{P_{IP3}} \right)^2 \)

from our previous analysis, can work out SFDR, etc.

• Basic approach: consider IMR when \( P_{in} = P_{max} \)

\[
IMR = \frac{P_{im}}{P_d} = \left( \frac{P_{max}}{P_{IP3}} \right)^2 \\
P_{im} = P_{min} \cdot G; \quad P_d = P_{max} \cdot G \\
IMR = \frac{P_{min} G}{P_{max} G} = \frac{P_{min}}{P_{max}} = \left( \frac{P_{max}}{P_{IP3}} \right)^2
\]
Spur Free Dynamic Range

• Since \( IMR = \frac{P_{\text{min}} G}{P_{\text{max}} G} = \frac{P_{\text{min}}}{P_{\text{max}}} = \left( \frac{P_{\text{max}}}{P_{IP3}} \right)^2 \)

it follows (just re-arranging) that:

\[ P_{\text{min}} = \frac{P_{\text{max}}^3}{P_{IP3}^2}; \quad P_{\text{max}} = P_{\text{min}}^{1/3} \cdot P_{IP3}^{2/3} \]

• Remember:
  – \( P_{\text{min}} \) comes from noise analysis, so is known
  – \( \text{SFDR} = \frac{P_{\text{max}}}{P_{\text{min}}} \)

• Final result:

\[ \text{SFDR} = \left( \frac{P_{IP3}}{P_{\text{min}}} \right)^{2/3} \]
Spur Free Dynamic Range

- Careful: previous page was all in MKS (or similar) units
- Usually specify these things in dB:

\[
SFDR = \left( \frac{P_{IP3}}{P_{\text{min}}} \right)^{2/3}
\]

Becomes:

\[
SFDR(dB) = \frac{2}{3} \left[ P_{IP3}(dB) - P_{\text{min}}(dB) \right]
\]

- Not complicated—just be careful
- One final note: the book gets the same results, but from another path; Pozar does the analysis from the point of view of the output power (vs. input like done here).
Congratulations

• You made it to the end.

• We’ll have a review session at our regular class period on Thursday – come with your questions about the final exam
Topics Covered This Year

- RF models – lumped element RLC models of common components (l< $\lambda/10$)
  - Behavior, resonances (series, parallel), some uses of these parasitic phenomena
- Electromagnetic analysis of transmission line structures
  - Maxwell’s eq., boundary conditions, assumptions made for solution
- Transmission-line models
  - Derivation from lumped-element sections, parameters ($\alpha$, $\beta$, $\gamma$), significance of each
  - Meaning and role of $Z_o$, $v_p$, $\lambda$, etc.
  - Source, load mismatch effects; reflection coefficients, impedance transforms, VSWR, translation along lines, boundary conditions
  - Power transmission
  - Stub impedance/admittance; origin, uses
- Fun with Smith charts
- Lumped-element matching network design
- Distributed circuits
  - $\lambda/4$ transformers
  - Series-line matching networks, single-stub matching networks, double-stub
  - Limitations, design procedures, detailed understanding
  - Bandwidth effects in matching networks
- Network analysis
  - $S$, $Y$, $Z$ parameters – matrix representations, definitions, finding matrix elements from circuits
  - Circuit analysis using matrix representations
  - Flow graphs, Mason’s rule
  - Generalized s-parameters
Topics Covered This Year (2)

• Amplifier design cases as considerations
  – Simultaneous conjugate matching conditions – maximum gain
  – Design for specified gain; gain circles, trade-offs (e.g. for bandwidth)
  – Design for noise figure; noise figure circles, interaction with gain circles
  – Detailed understanding and ability to design circuits

• Power gain definitions & use
  – Operating power gain ($G_P$), available power gain ($G_A$), transducer power gain ($G_T$)
  – Definitions, significance of each
  – What is each good for?

• Stability
  – Meaning/significance of stability
  – Source, load stability circles
  – Interpretation of the circles
  – $k$-$\Delta$, $\mu$ tests and what they mean
  – Unconditional stability: definition, concepts

• Nonlinear effects
  – Derivations, foundations
  – Gain compression
  – Intermodulation
  – Definitions: $P_{IP3}$, $P_{1dB}$, IMR, blocking, desensitization, cross modulation
  – Input/output spectra, frequencies present, etc.

• Noise
  – Thermal noise in resistors/passives
  – Available noise
  – Effective noise temperature, noise resistance
  – Noise in 2-ports: $T_E$, $F$, NF
  – Noise measurement ($Y$ factor)
  – Cascaded noise figure
  – Receiver sensitivity (MDS)
  – Dynamic range (SFDR)