

Case Study Competition 2008

# Go Wireless!

Testing for future communications standards

**Final Round Exercises**

**75** Years of  
Driving  
Innovation

 **ROHDE & SCHWARZ**

# Scenario

You are an RF development engineer employed by a manufacturer of wireless communications equipment, where you are a member of a team that is supposed to develop devices for the Long-Term Evolution (LTE) standard, a potential successor to UMTS.

This case study provides the terms of reference for the characterization of various components and the design of individual modules. As a versatile and highly regarded engineering professional, your skills are also very much in demand in related departments. In addition to your specialty – RF technology – you have a good reputation in the area of baseband coding.

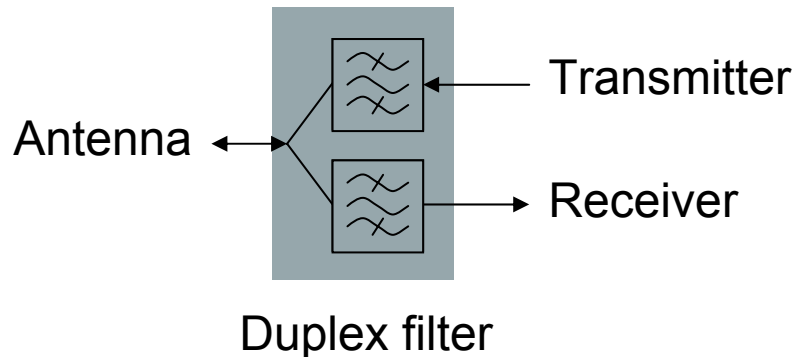


# Exercise I: Duplex Filter and Antenna

## Task description

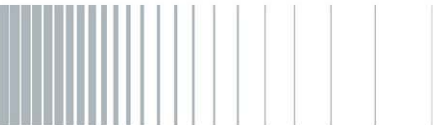
For a next-generation mobile phone, you have to characterize a duplex filter. Furthermore, your task is to dimension a circuit that matches the antenna via this filter to the output amplifier of a particular transmit band.

The job of a duplex filter is to separate the transmit path (uplink) from the receive path (downlink) in a frequency band. Surface acoustic wave (SAW) filters are normally used for this purpose.



# Duplex Filter

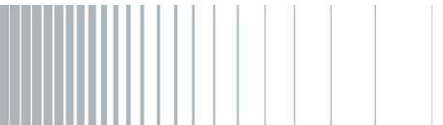
Nowadays RF transmission lines in mobile phones as well as the microwave integrated circuits (MMICs) used in these phones are increasingly being implemented using balanced lines instead of conventional  $50\ \Omega$  single-ended technology. When balanced lines are used, the duplex filter usually has a second function in addition to acting as a frequency splitter: It serves as a balanced to unbalanced (balun) transformer in order to match the unbalanced antenna to the balanced transmission line.



# Duplex Filter

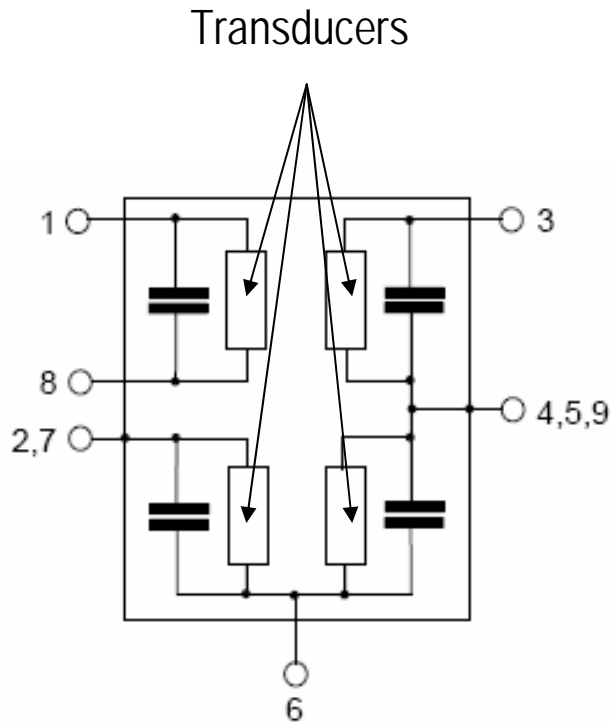
## Exercise I.A.1 (4 points)

- I Which wave modes can propagate on a balanced transmission line?
- I Which mode is used to transfer the desired signal?
- I What advantages do balanced lines offer relative to unbalanced lines?



# Duplex Filter

Example: Duplex filter with an single-ended (ground-referenced) port for the transmitter output and a balanced port for the receiver input.

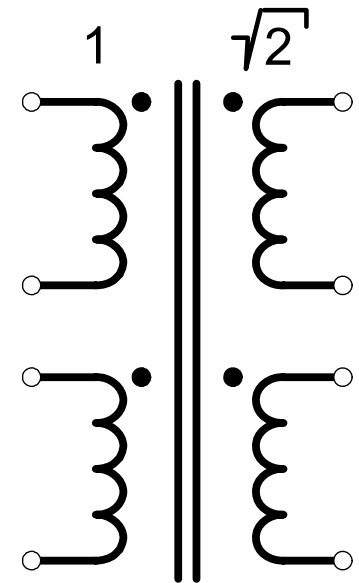


## Pin configuration

|               |                |
|---------------|----------------|
| 3             | se. TX Input   |
| 1,8           | bal. RX Output |
| 6             | Antenna        |
| 2, 4, 5, 7, 9 | Ground         |

# Duplex Filter

You want to test this duplex filter for compliance with specifications. It is soldered to an evaluation board with female coaxial connectors. The effect of the board can be assumed to be negligible for the purpose of the test to be performed. You have a vector network analyzer (VNA) with four single-ended measuring ports that you can use. In order to eliminate systematic errors and to allow for exact scattering parameter measurements using the VNA, you have calibrated it at the coaxial connectors of the connected measuring cables. You also have two hybrid transformers available with a transfer ratio of  $1:\sqrt{2}$ . They can be regarded as ideal transformers.



Schematic diagram of a hybrid transformer

# Duplex Filter

## Exercise I.A.2 (7 points)

Draw the block diagram of the test setup for the following port arrangement:

| VNA       | Filter port | Mode              | Reference impedance |
|-----------|-------------|-------------------|---------------------|
| Test port |             |                   |                     |
| 1         | Antenna     | Single-ended      | 50 $\Omega$         |
| 2         | Tx Input    | Single-ended      | 50 $\Omega$         |
| 3         | Rx Output   | Differential mode | 100 $\Omega$        |
| 4         | Rx Output   | Common mode       | 25 $\Omega$         |

# Duplex Filter

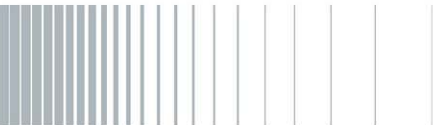
## Exercise I.A.3 (2 points)

You find that the measured value of the stopband attenuation of the duplex filter between the transmitter output and the receiver input in the uplink channel is limited by the internal noise of the VNA receiver. This means that the measured value is only 43 dB instead of 50 dB as specified in the data sheet. The measuring bandwidth of the VNA can be set to a value between 1 Hz and 10 MHz in a 1-2-5 sequence. You have presently configured the VNA for a measuring bandwidth of 1 MHz. You must reduce the measuring bandwidth in order to measure the stopband attenuation of the filter correctly. What is the maximum bandwidth that you can use for this purpose? Explain your answer.

# Duplex Filter

Note: you can assume that with a correct measurement the dynamic range of the VNA, which is limited by white noise of the receiver, is at least 6 dB larger than the specified stopband attenuation of the filter.

What other setting of the VNA could you change instead, and by how much, in order to achieve the dynamic range necessary for the measurement?



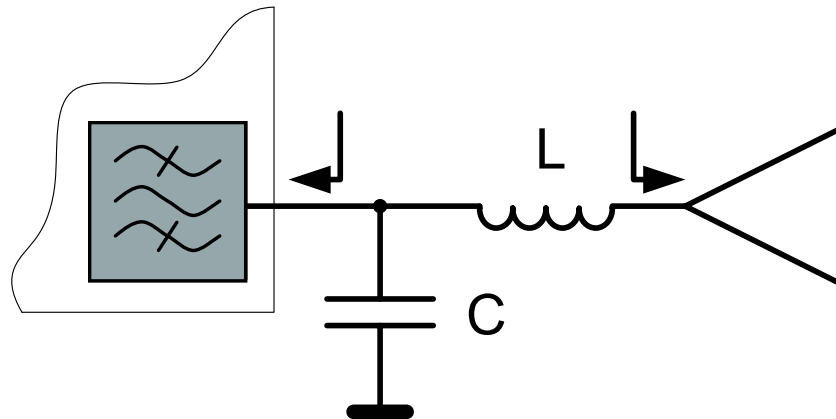
# Antenna Reflection Coefficient

If you know the common-mode and differential-mode input impedances of the receiver, you can describe the transfer characteristics of the duplex filter in the single path transmitter output – antenna by means of a two-port scattering matrix  $S_{TA}$ . With a modern VNA, you can import an externally calculated  $S_{TA}$  matrix and take it virtually into account while measuring the antenna reflection coefficient. You can use this computational step, which is called “embedding”, to obtain the transmitter port input reflection coefficient of the duplex filter with the antenna, even though you are only making measurements on the antenna connected to the VNA.



# Antenna Reflection Coefficient

At 1950 MHz, the midband frequency of the uplink channel for LTE Operating Band I, the complex reflection coefficient of the transmitter input port of the duplex filter (referenced to  $50 \Omega$ ) is  $\rho_D = -0.083 - j0.275$ . Specify the component values of an LC network (see drawing) that will match this load reflection at 1950 MHz to the impedance of the output stage of the transmit branch, which is  $Z_A = (10 - j5) \Omega$ .



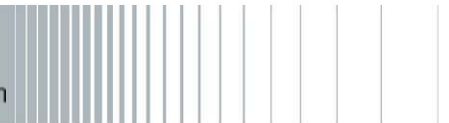
# Antenna Reflection Coefficient

## Exercise I.B.1 (9 points)

Determine the values of the matching network components L and C. You can either calculate the values or determine them graphically using the Smith chart.

## Exercise I.B.2 (3 points)

In order to reduce costs, you are supposed to design a circuit that uses two unbalanced transmission lines, which can be assumed to be ideal  $50 \Omega$  lines, to provide the desired matching without using discrete components. Describe a suitable circuit in qualitative terms.



# Exercise II: Synthesizer Core

## Task description

Your task is to develop a new synthesizer core for the next-generation base station. In addition to the latest wireless communications standards, which are described under the name 3GPP-LTE, future base stations must also be able to support older standards such as PCS 1900. It must be possible to use the synthesizer for both paths (transmit and receive) in order to save development time. In order to be able to meet the very high system linearity requirements for modern modulation methods, your project manager has decided to use a direct conversion concept.

The equipment will be used outdoors, such as on rooftops, so the required operating temperature range is  $-35\text{ }^{\circ}\text{C}$  to  $+75\text{ }^{\circ}\text{C}$  ( $-30\text{ }^{\circ}\text{F}$  to  $+170\text{ }^{\circ}\text{F}$ ).

# Requirements from the Standard

## Available frequency bands

Table 1: Frequency bands in accordance with 3GPP-LTE (FDD)

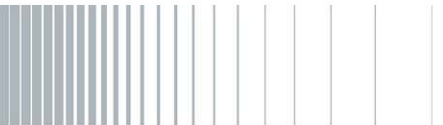
| Operating band | UL frequencies<br>UE transmit, Node B receive | DL frequencies<br>UE receive, Node B transmit |
|----------------|---|---|
| I              | 1920 MHz to 1980 MHz                          | 2110 MHz to 2170 MHz                          |
| II             | 1850 MHz to 1910 MHz                          | 1930 MHz to 1990 MHz                          |
| III            | 1710 MHz to 1785 MHz                          | 1805 MHz to 1880 MHz                          |
| IV             | 1710 MHz to 1755 MHz                          | 2110 MHz to 2155 MHz                          |
| V              | 824 MHz to 849 MHz                            | 869 MHz to 894 MHz                            |
| VI             | 830 MHz to 840 MHz                            | 875 MHz to 885 MHz                            |
| VII            | 2500 MHz to 2570 MHz                          | 2620 MHz to 2690 MHz                          |
| VIII           | 880 MHz to 915 MHz                            | 925 MHz to 960 MHz                            |
| IX             | 1749.9 MHz to 1784.9 MHz                      | 1844.9 MHz to 1879.9 MHz                      |
| X              | 1710 MHz to 1770 MHz                          | 2110 MHz to 2170 MHz                          |
| XI             | 1427.9 MHz to 1452.9 MHz                      | 1475.9 MHz to 1500.9 MHz                      |
| XII            | 698 MHz to 716 MHz                            | 728 MHz to 746 MHz                            |
| XIII           | 777 MHz to 787 MHz                            | 746 MHz to 756 MHz                            |
| XIV            | 788 MHz to 798 MHz                            | 758 MHz to 768 MHz                            |

# Requirements from the Standard

## Frequency stability

Table 2: Required frequency stability

| <b>Time</b>             | <b>Accuracy</b> |
|-------------------------|-----------------|
| <b>within time slot</b> | $\pm 0.05$ ppm  |
| <b>within 1 day</b>     | $\pm 0.2$ ppm   |
| <b>within 1 month</b>   | $\pm 0.5$ ppm   |



# Requirements from the Standard

## Noise immunity (PCS 1900)

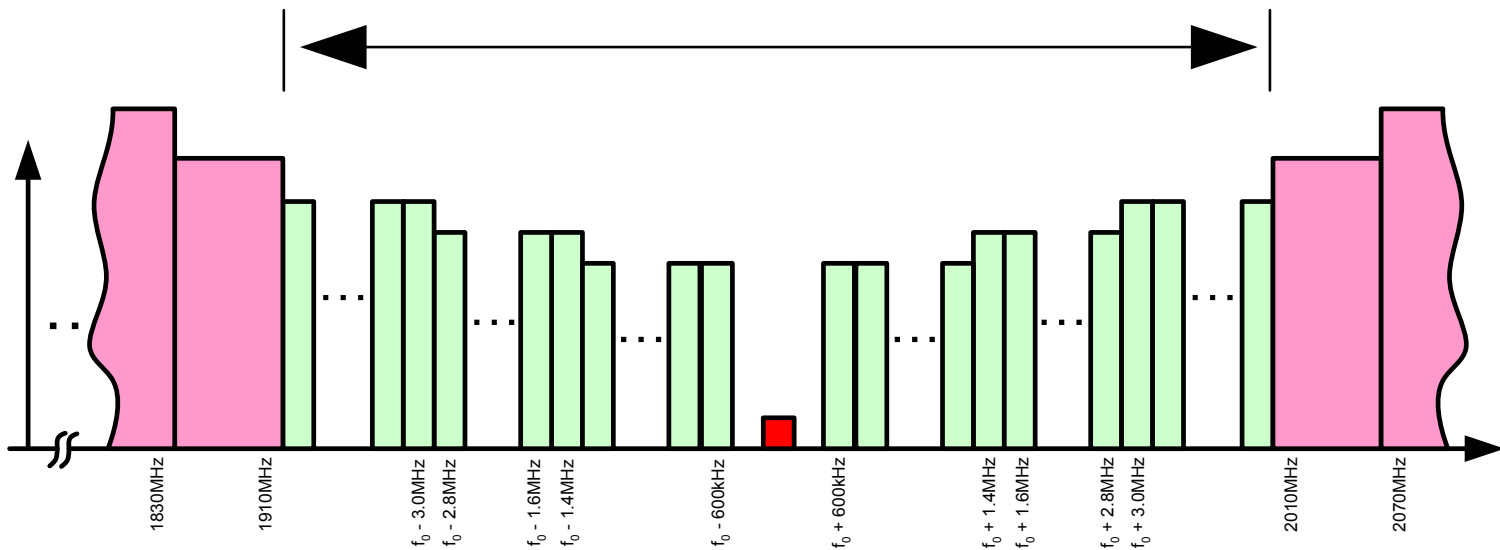


Figure 1: Blocking profile for PCS1900

Red: desired signal / Nutzsignal

Green: possible interference signal / möglicher Störer

Light red: interference signal out of band /  
Störer außerhalb des Mobilfunkbandes

# PLL Design

## Exercise Part II.A

After an initial study of the requirements, you have decided to use a single-loop PLL (phase-locked loop) design. The block diagram looks like this:

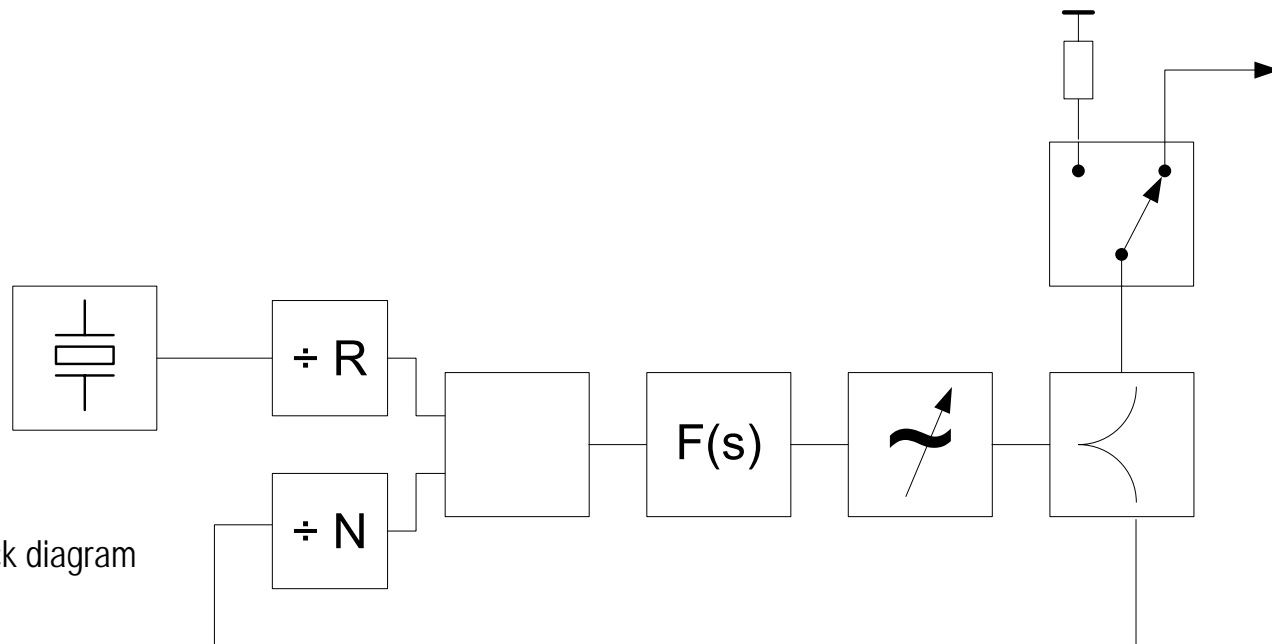


Figure 2: PLL block diagram

# PLL Design

## Exercise II.A.1 (3 points)

The normal channel spacing  $f_{ch}$  in the wireless communications bands is 200 kHz, but there are some regions in which a 100 kHz spacing is used. This means it must be possible to set the output frequency with a spacing of 100 kHz over the range of 698 MHz to 2690 MHz. Calculate the values of the PLL division factors R and N under the assumption that the PLL is an integer-N type and the reference crystal oscillates at a frequency of 20 MHz.



# PLL Design

## Exercise II.A.2 (5 points)

During your simulations, you find that you have to design the loop with a control bandwidth of around 10 kHz so the circuit will have low transient overshoot, be stable in all operational conditions, and deliver a signal with low spurious content. Unfortunately, you cannot achieve the desired channel-switching times. Would it help if you could use a fractional-N divider in the feedback path, and what disadvantages would this have? Briefly describe the changes to the synthesizer core that would be necessary.



# PLL Design

## Exercise II.A.3 (2 points)

Your supplier can provide AT-cut and SC-cut quartz crystals for the reference crystal. Excerpts from the data sheets are shown on the following pages. Is the temperature stability of either of these crystals sufficient to allow the specified frequency stability to be achieved?

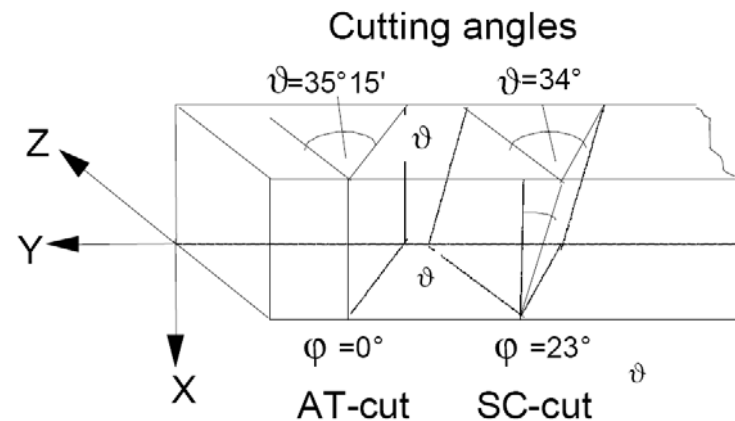


Figure 3: Quartz crystal cutting variants

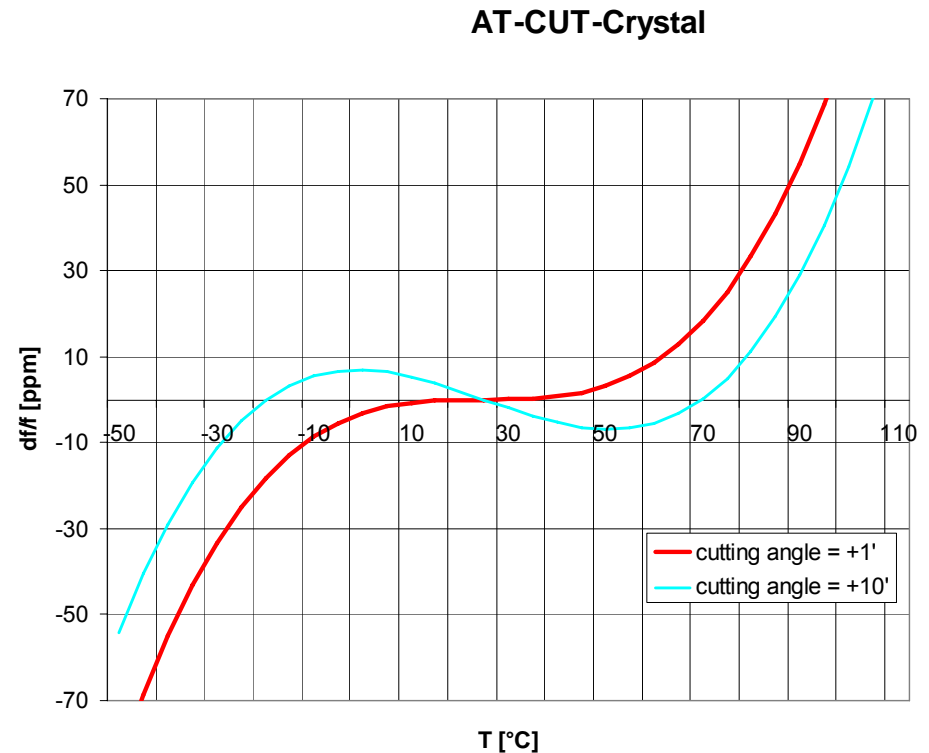
# Data Sheet for Crystal 1 (AT Cut)

AT-cut quartz crystal

Frequency: 20.0 MHz  
Adjustment tol.:  $\pm 3$  ppm  
Aging/year:  $2 \times 10^{-7}$   
Short-term stability:  $5 \times 10^{-10} / 10$  s  
Quality factor:  $1 \times 10^6$

Temp. stability: see diagram

Application: for use at room temp.



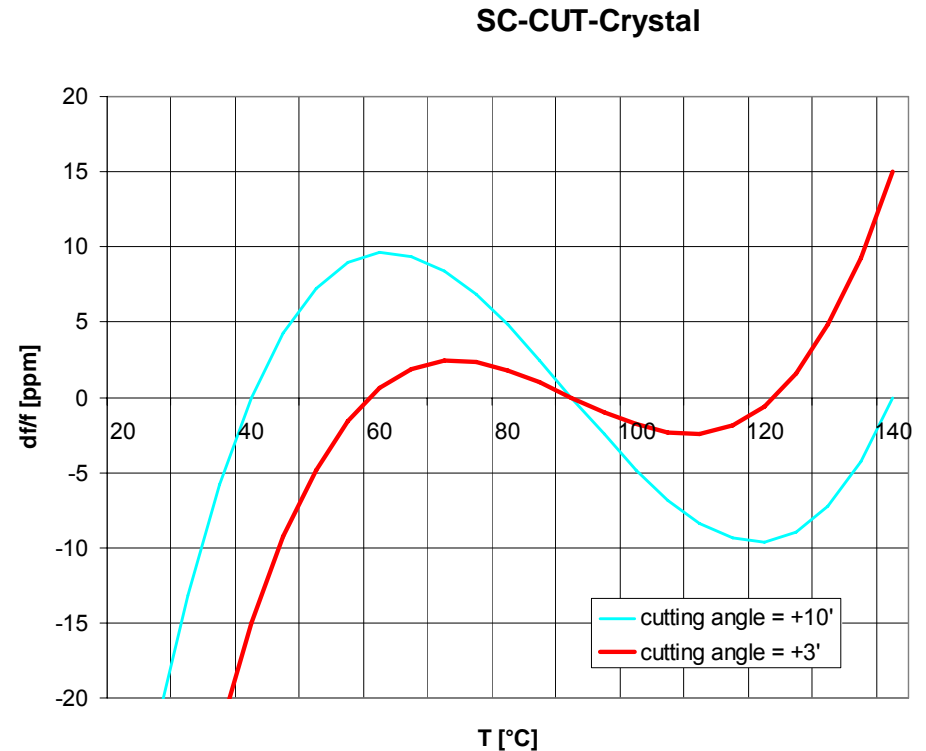
# Data Sheet for Crystal 2 (SC Cut)

SC-cut quartz crystal

Frequency: 20.0 MHz  
Adjustment tol.:  $\pm 1$  ppm  
Aging/year:  $5 \times 10^{-8}$   
Short-term stability:  $2 \times 10^{-11} / 10$  s  
Quality factor:  $1 \times 10^6$

Temp. stability: see diagram

Application: for use at high temp.

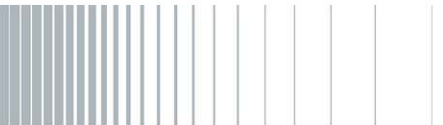


# PLL Design

## Exercise II.A.4 (5 points)

Consider measures that could be used to enable the requirements to be met with one of the two crystals at minimal expense (about € 10 / US\$ 15).

Which of the two crystals would you use? Explain your choice.



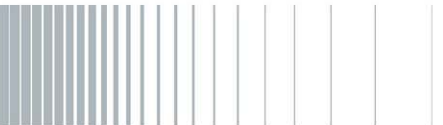
# Phase Noise

## Exercise Part II.B

Unfortunately, synthesizers do not generate perfectly clean output signals. The output signal fluctuates slightly in amplitude and frequency. The frequency fluctuation can also be regarded as phase fluctuation, in the same way that frequency modulation and phase modulation are related. For this reason, synthesizers are said to have phase noise (PN). It is specified in units of dBc/Hz, which means that the noise power is referenced to a bandwidth of 1 Hz.

As a consequence of this phase noise, frequency ranges outside the actual LO frequency are downconverted into the desired signal band, which degrades the signal to noise ratio (S/N) of the desired signal.

This problem is illustrated by the following diagram:



# Phase Noise

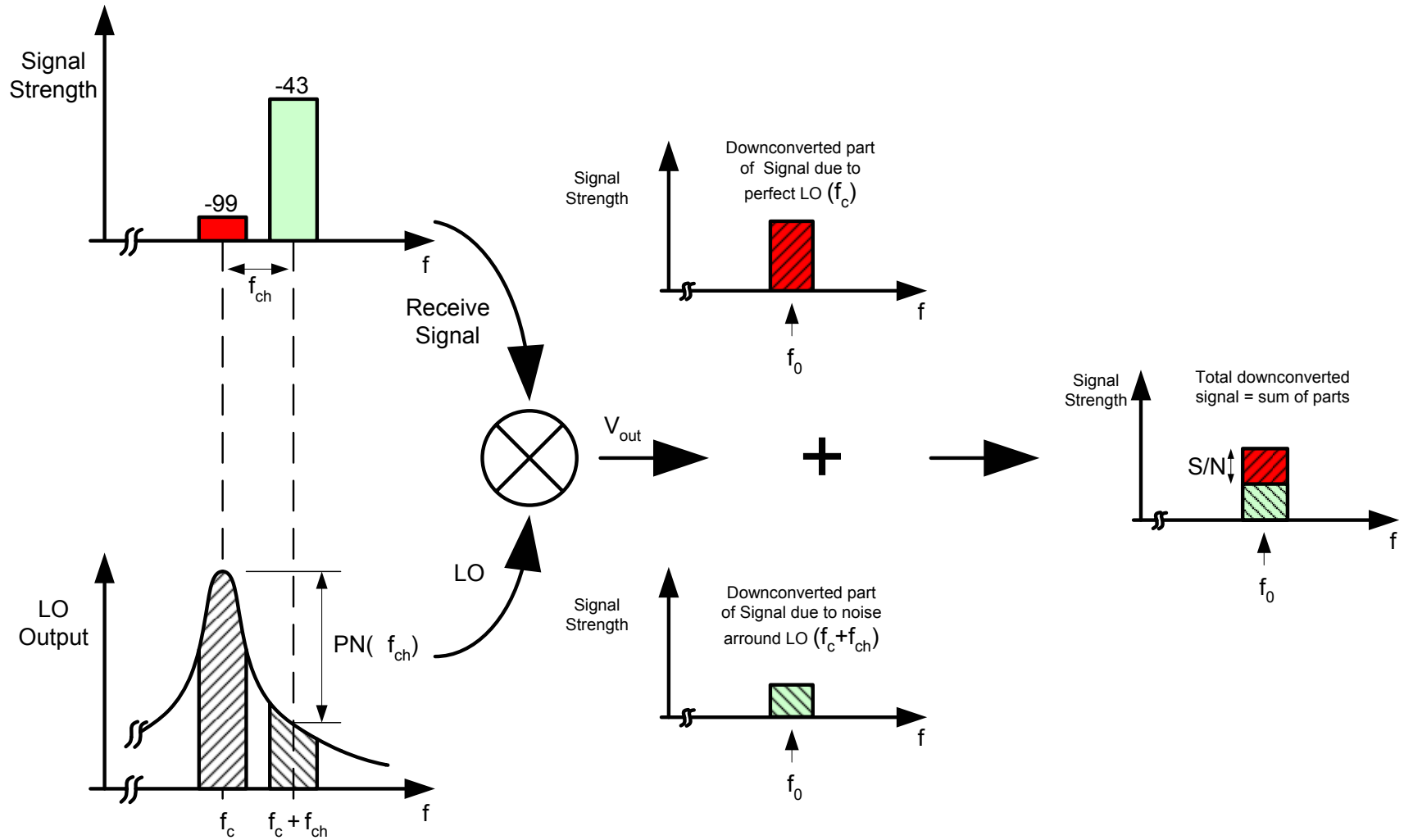


Figure 4: Reciprocal mixing of phase noise

# Phase Noise

## Exercise II.B.1 (7 points)

Calculate the maximum amount of phase noise the synthesizer can have without causing any problems for PCS 1900 (see Figure 1) due to the previously described effect. For this purpose, assume that a minimum signal to noise ratio (S/N) of 9 dB is required for PCS 1900.

**Note:** As the specified value of the phase noise is referenced to a bandwidth of 1 Hz (dBc/Hz), while the channel bandwidth is 200 kHz with PCS 1900, you must take the following relationship into account in your power calculation:

$$C/N = PN + 10 \log(200 \text{ kHz})$$

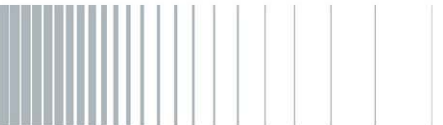
# Phase Noise

## Exercise II.B.2 (3 points)

What is the relationship between the phase noise of the reference crystal and the phase noise of the full synthesizer?

Does the broadband noise of the reference crystal also affect the broadband noise of the full synthesizer?

Take the results from Part A into account.

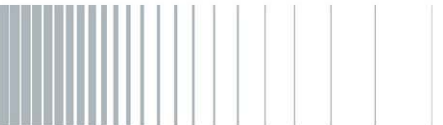


# Exercise III: Power Amplifier Production Test

## Task description

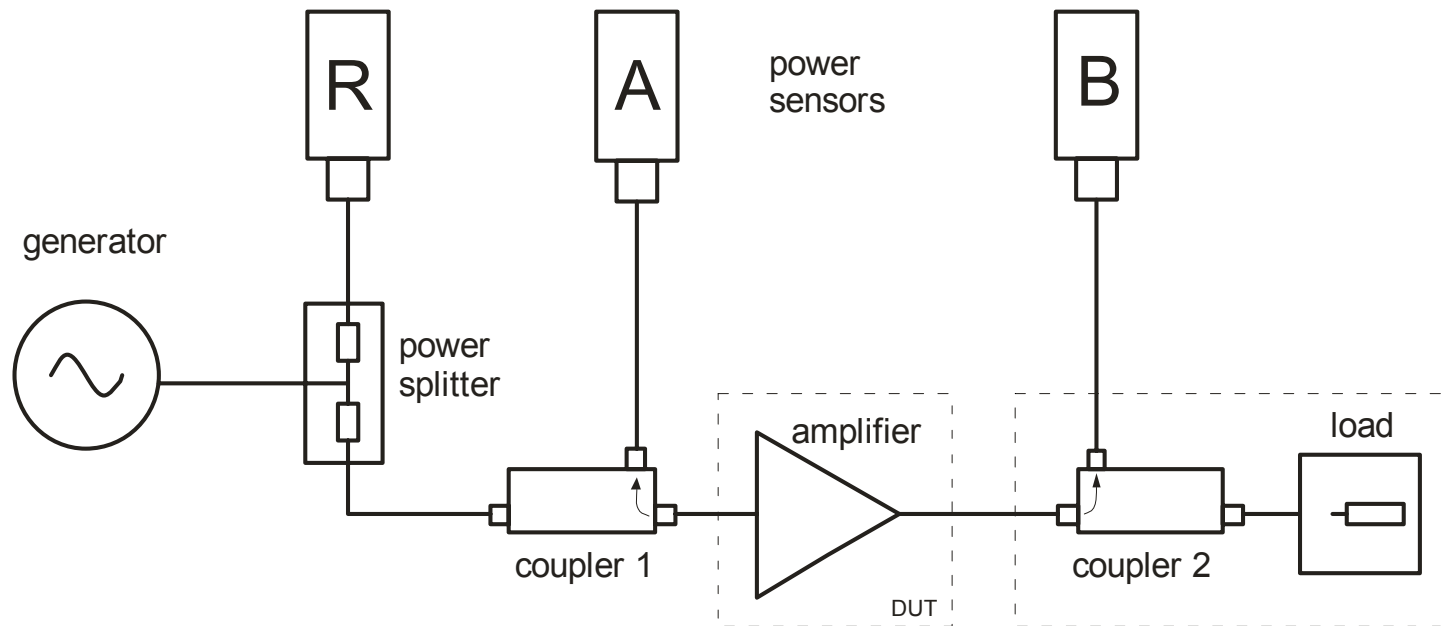
A related department asks you to assist in the design of a production setup for an LTE base station. This includes the characterization of the RF power amplifier in terms of its input reflection coefficient, gain and output power. It is essential to know the values of these parameters, as well as several others, in order to assure reliable operation.

The key aspect of this exercise is analyzing the measurement uncertainty. The results will be needed for selecting the components that determine the accuracy of the design, and you will decide whether the system design is adequate.



# Test Setup

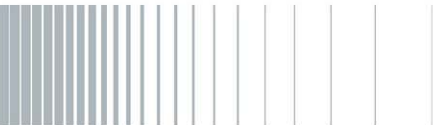
The following scalar test setup should be used for the characterization:



# Test Setup

A signal generator with high spectral purity is used as the signal source. Its output signal is fed via a power splitter to the device under test and a power sensor used as a reference. A directional coupler is inserted in the signal path at the amplifier input. It couples out part of the reflected wave for measurement using a power sensor.

The output signal of the amplifier is fed to a dummy load, with part of the output wave being coupled out to another power sensor (B) for measurement.



# System Equations

## Exercise Part III.A (5 points)

For this test setup, state how the input reflection coefficient  $|S_{11}|$ , gain coefficient  $|S_{21}|$  and output power  $P$  of the amplifier can be determined from the values measured using the power sensors and the coupling coefficients of directional couplers 1 and 2.

In your equations, ignore all other parameters that affect the measurement, such as the asymmetry of the power splitter, impedance mismatch between pairs of components, insertion loss of the directional couplers and the measurement uncertainty of the power sensors.

# Input Reflection Coefficient

## Exercise III.B (10 points)

With scalar measurement of the input reflection coefficient, the directivity of the directional coupler used for making the measurement determines the accuracy of the measured value. Describe this effect in quantitative terms and answer the following question: What is the minimum directivity that directional coupler 1 must have in order to ensure that amplifiers with an input  $VSWR \geq 2.0$  can be recognized as rejects with full certainty and all amplifiers with a  $VSWR \leq 1.7$  will pass the test?

Note: The stated numerical values should refer to the actual VSWR values and not the measured values. As with Exercise Part A, all other parameters that could affect the results are to be ignored.

# Output Power

## Exercise Part III.C

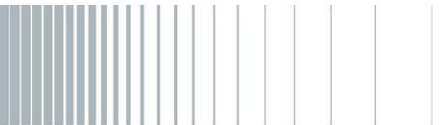
Power amplifiers in base stations are subjected to stringent output power requirements (into  $50 \Omega$ ). For this reason, it is necessary to determine the output power as accurately as possible, or at least to know the magnitude of the measurement uncertainty. The following parameters must be taken into account for the test setup described here:

- Impedance mismatch of the amplifier output
- Uncertainty of the coupling coefficient of directional coupler 2
- Measurement uncertainty of power sensor B

The following information in this regard is available from measurements and data sheet specifications:

# Output Power

- The output reflection coefficient of the amplifier is characterized by a VSWR of 1.6, and the phase of the reflection coefficient can be assumed to be distributed uniformly.
- The measured value of the magnitude of the reflection coefficient at the input of the directional coupler is 0.051 with a dummy load connected.
- The measured value of the coupling attenuation is  $(30.06 \pm 0.07)$  dB.
- A value of  $\pm 0.081$  dB can be assumed for the worst-case measurement error of the power sensor.



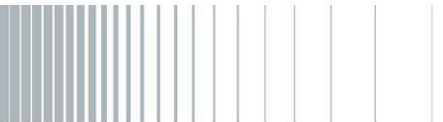
# Output Power

## Exercise III.C.1 (7 points)

Due to mismatch of the dummy load, the power supplied by the amplifier during testing is different from what it would supply with a perfectly matched load ( $50 \Omega$ ). Calculate the potential maximum error in percent and dB, and incorporate the result into the appropriate system equation in Exercise Part A.

## Exercise III.C.2 (3 points)

Calculate the worst-case measurement error of the (indirectly) measured amplifier output power compared to the true value (into  $50 \Omega$ ). Take the three parameters mentioned in the task description into account in your calculations.

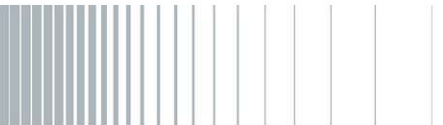


# Exercise IV: MIMO Baseband Coding

## Task description

You have been asked to assist the baseband processing team of your mobile equipment group in their task to find a way to maximize the downlink data rate experienced by the end user. To achieve this, multiple antennas can be used in the base station (eNodeB) as well as in your user equipment. The physical distance between the UE antennas can be as much as several wavelengths, since your equipment is designed to be used in a laptop computer environment.

To achieve your design goals, you analyze the theoretical rate limit (Shannon capacity) of your MIMO channel in the first step, and then design transmitter and receiver channel matrices in the second step that can be used in a real-life environment.



# Capacity of Single-User MIMO

## Exercise Part IV.A:

Given  $\underline{y}$  a received vector of  $N$  elements, where  $N$  corresponds to the number of receiver antennas in a single-user MIMO system:

$$\underline{y} = \underline{x} + \underline{n}$$

where  $\underline{x}$  is the signal part and  $\underline{n}$  is the white Gaussian noise. It is assumed that all signals have zero mean and are Gaussian distributed.

The channel capacity  $C$  [bit/Hz] can be determined with

$$C = \frac{1}{2} \sum_{i=1}^N \log_2 \left( 1 + \frac{\sigma_{x_i}^2}{\sigma_n^2} \right)$$

assuming for Gaussianity, uncorrelated output streams ( $\underline{y}$ ), which is equivalent to independence.

$\sigma_{x_i}^2$  is the mean power of the received signal part of antenna  $i$ .

$\sigma_n^2$  is the mean noise power.

# Channel Capacity

## Exercise IV.A.1 (1 point)

Now assume that a total amount of power  $P_T$  can be distributed equally on all antennas of  $\underline{x}$  (only realistic if the channel matrix  $H$  is a diagonal identity matrix).

Assuming  $\frac{P_T}{\sigma_n^2} = 1000$ , i.e. 30 dB, what is the maximum achievable rate [bit/Hz] if one antenna is used?

## Exercise IV.A.2 (1 point)

How much do we get if we use two antennas with equal power spread

$$\sigma_{x_1}^2 = \sigma_{x_2}^2 = \frac{P_T}{2} ?$$

# Channel Capacity

## Exercise IV.A.3 (2 points)

How much capacity gain would you expect between the 2 previous scenarios when  $\frac{P_T}{\sigma_n^2} \rightarrow \infty$  ?

## Exercise IV.A.4 (1 point)

Now try with  $N = 4$  antennas and assuming

$$\sigma_{x_1}^2 = \sigma_{x_2}^2 = \sigma_{x_3}^2 = \sigma_{x_4}^2 = \frac{P_T}{4}$$

How much would be the achievable rate if  $\frac{P_T}{\sigma_n^2} = 30\text{dB}$  and what is the

asymptotic gain ( $\frac{P_T}{\sigma_n^2} \rightarrow \infty$ ) compared to the case of 1 antenna?

# Channel Capacity

## Exercise IV.A.5 (2 points)

What would the achievable raw data rate [bit/s] be in this case (four antennas) if we have an LTE signal fitting into a channel of  $BW = 20$  MHz, OFDM FFT (2048 carriers, of which 1200 can be used for data) and an FFT sample rate of  $f_s = 30.72$  MHz? ( $\frac{P_T}{\sigma_n^2} = 30$  dB)

## Exercise IV.A.6 (1 point)

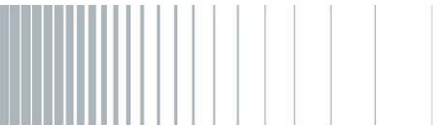
Name two available modulations in LTE downlink.

# Channel Capacity

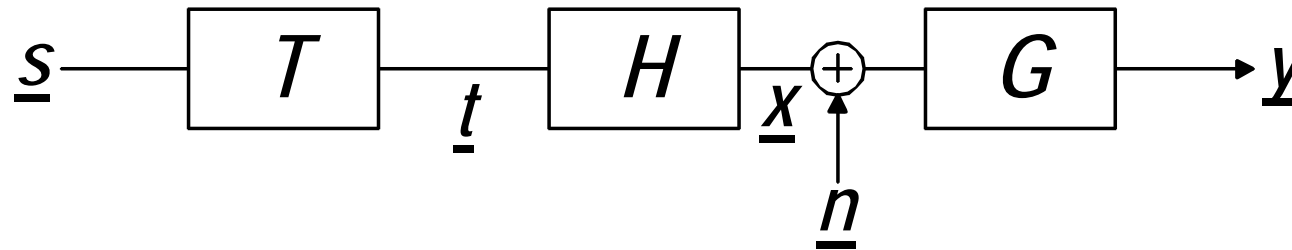
## Exercise IV.A.7 (1 point)

Up to now, we have analyzed the theoretically achievable maximum rates under the assumption of Gaussian distribution. In real-life implementations, only quantized constellations can be used.

What is the influence of having quantized constellations instead of Gaussian continuous distribution?



# Transmit & Receive Matrix Design for Downlink LTE MIMO



$\underline{s}$ : Input vector of size  $s$ , where  $s$  corresponds to the number of layers.  $s$  is assumed to be the number of input streams, transmit antennas and receive antennas.

Assume all symbols are uncorrelated.  $T$  is the transmit matrix to be designed in order to maximize the data rate that can be achieved, assuming zero mean Gaussianity. There is a transmit power constraint:

$$\mathbf{E}\{\underline{t}^H \underline{t}\} = P_T = \sum P_i$$

$\underline{t}$  represents the transmit vector,  $\underline{t}^H$  is the Hermitian of  $\underline{t}$ ,  $\mathbf{E}\{\underline{t}^H \underline{t}\}$  is the expectation of  $\underline{t}^H \underline{t}$ .

# Matrix Design

## Exercise Part IV.B

The single-value decomposition (SVD) of the channel matrix  $H$  is defined as

$$\mathbf{H} = \mathbf{U}\mathbf{D}\mathbf{V}^H$$

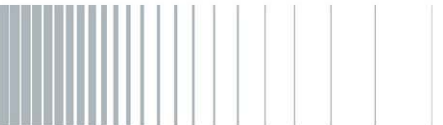
To diagonalize the output correlation matrix, we propose the following transmit matrix:

$$\mathbf{T} = \mathbf{V}\mathbf{P}^{1/2}$$

and the following receive matrix:

$$\mathbf{G} = \mathbf{U}^H$$

$\mathbf{P}$  is the diagonal matrix of elements  $P_i$ .  $P_i$  is the power allocated to antenna  $i$ .



# Matrix Design

The maximum achievable rate [bit/Hz] is

$$C = \frac{1}{2} \sum_{i=1}^s \log_2 \left( 1 + \frac{P_i}{\sigma_n^2} \lambda_i \right)$$

$\lambda_j$ : eigen values of diagonal matrix  $D$ .

and 
$$\sum_{i=1}^s P_i = P_T$$

$P_i$  needs to be chosen such that  $C$  is maximized while observing the power constraint. This can be achieved by applying Lagrange multiplier methods to the above formula. The result is:

$$P_i = P_T \left( \frac{1}{s} + \left( \frac{\sigma_n^2}{P_T} \right) \left( \frac{1}{s} \sum_{j=1}^s \frac{1}{\lambda_j} - \frac{1}{\lambda_i} \right) \right)$$

# Matrix Design

## Exercise IV.B.1 (5 points)

Assume a 2x2 MIMO channel with

$$\mathbf{H} = \begin{bmatrix} 0.98995 & 0.98995 \\ 0.56569 & -0.56569 \end{bmatrix}$$

$$s = 2$$

LTE defines the following codebooks for 2x2 MIMO:

$$0: \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \quad 1: \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & 1 \\ 1 & -1 \end{bmatrix} \quad 2: \frac{1}{\sqrt{2}} \begin{bmatrix} 1 & 1 \\ j & -j \end{bmatrix}$$

Compute the SVD of  $\mathbf{H}$ .

# Matrix Design

## Exercise IV.B.2 (1 point)

Determine  $\lambda_1$  and  $\lambda_2$ .

## Exercise IV.B.3 (1 point)

Find the optimal power allocation assuming  $\frac{P_T}{\sigma_n^2} = 30$  dB

## Exercise IV.B.4 (1 point)

What is the effect of high  $\frac{P_T}{\sigma_n^2}$  on the power allocations?

## Exercise IV.B.5 (1 point)

Determine the maximum achievable rate [bit/Hz] based on the power allocations.

# Matrix Design

## Exercise IV.B.6 (2 points)

Determine  $T$  and  $G$ , where power allocations observe the constraint of B4.

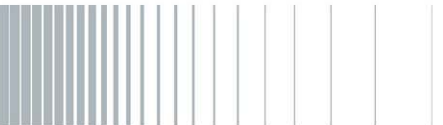
## Exercise IV.B.7 (2 points)

Which codebook number can be used as  $T$ ?

Tip: Phase offsets of any transmit antennas will not influence the capacity and thus optimality.

## Exercise IV.B.8 (3 points)

Please explain the advantage and disadvantage of using codebooks instead of user-configurable transmit matrices.



# On Your Mark, Get Set, Go!

We hope you enjoy solving the exercises presented here and that you arrive at the right results.

## A small comment:

The mathematics in the exercises is not difficult. The key to success is not the mathematics, but instead proper understanding of the situation.

