Design, realization and performance verification of high-frequency lock-in amplifiers for RGB-ITR

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Room: B.Brunelli – ENEA Frascati

Presentation Outline

1. Motivation and scopes
2. RGB-ITR measurements
3. RGB-ITR signals and system modelling
4. RGB-ITR simulator
5. Analog lock-in amplifier design
6. Lock-in performance verification
7. Questions without a final answer
8. Future developments
9. Conclusions

Motivations and Scopes::Status

Actual real-time signal detection, processing and data acquisition electronic section is realized using “**off-the-shelf**” standard laboratory equipment; therefore it is:

a. Cumbersome (1 m^3 approx.)

b. Heavy (>30 Kg: you need two guys to move the crates!)

c. Power hungry (200-500 Watt; you need a big PSU, not shown in the picture)

d. Expensive (10-15 kE/channel)

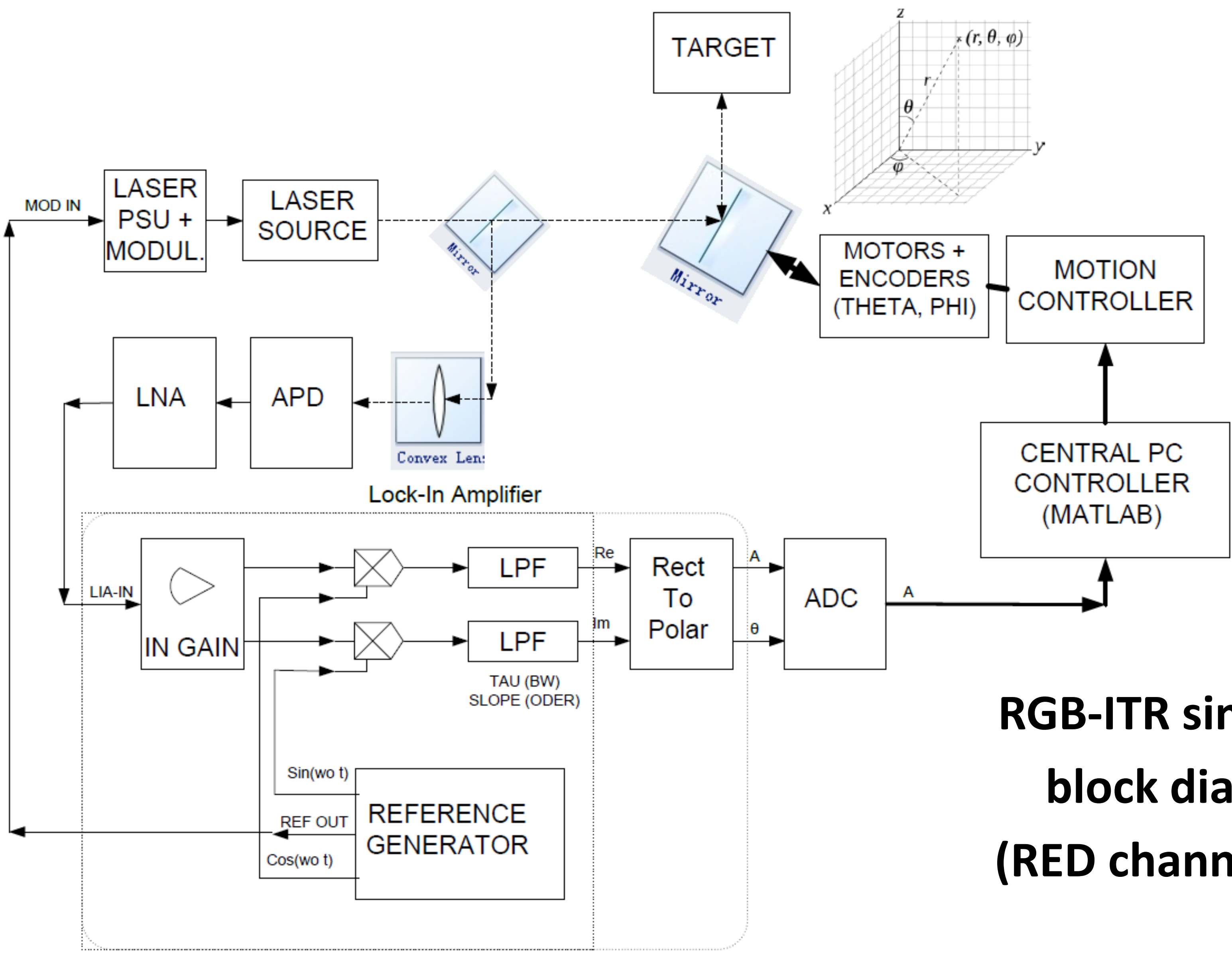
e. “Fragile” (not MIL-STD) hardware; not suitable for field deployment in “environmentally hard” (temperature, humidity, dust & dirt, etc.) and tight access places (e.g. catacombs).

Most of the actual limitations derive by the choice of commercial instruments designed for general purpose laboratory/bench usage *but not for field deployment*.

Q. #1: Are the instruments (**lock-in amplifiers**) used in the real-time signal processing section the right choice?

Q. #2: Is it possible **replace** the actual signal processing section with some (custom) electronic that avoid the previous five limitations?

RGB-ITR :: Imaging Topological RADAR



RGB-ITR simplified block diagram (RED channel only)

RGB-ITR electrical signal measurement parameters

Amplitude: amplitude (envelope) of the returning signal carry the information of the "color component" at the wavelength of the illuminating LASER spot on the target (RED channel: λ=660 nm, F_M=190 MHz, d=0.1 mm).

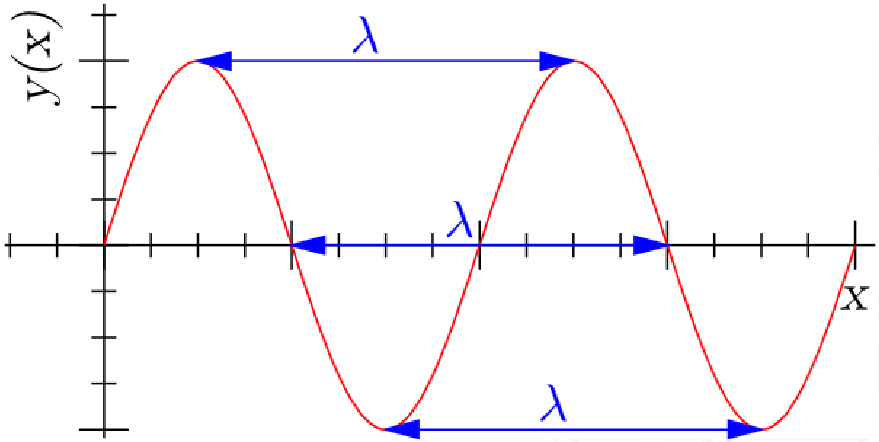
Phase: phase (relative to the reference modulating signal) carry the information of the distance of the target.

$$y(x,t) = A\cos(2\pi\frac{x}{\lambda} - \omega_M\, t) \;;\; \theta(d) = 2\pi\frac{x}{\lambda} \;;\; \lambda = \frac{c}{F_M} \;;\; \theta(d) = 2\pi x\frac{c}{F_M}$$

$$x = \frac{\theta}{2\pi}\,\frac{c}{F_M}\;; \quad r = 2x;$$

$$r = \theta\frac{c}{4\pi F_M}\;; \quad L = \frac{1}{2}\,\frac{c}{F_M}\;;$$

$$L_{RED} = 0.79m;\ \textit{(folding distance)}$$



Preliminary Assumptions

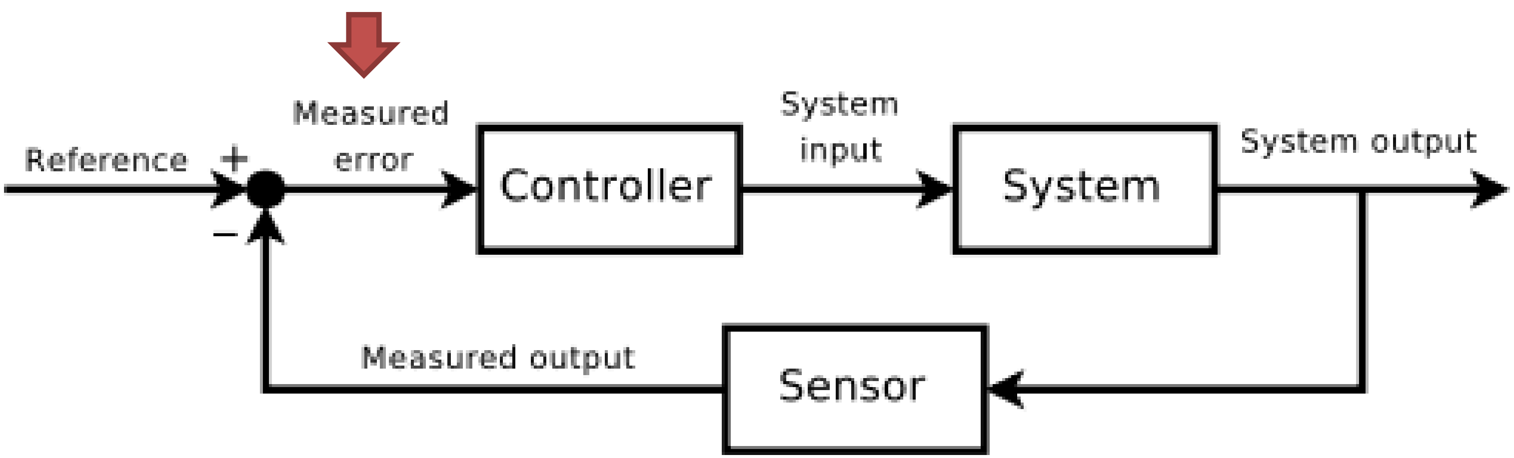
- We consider the optical components, the mirror motion control system, the optical paths and the specific proprieties of the target “out of scope”.
- Only the proprieties of the electrical signals are investigated using the available instruments: oscilloscopes (OS), spectrum analyzer (SA) and lock-in amplifier (LIA).
- Measurements are validated crossing time domain (OS) and frequency domain (SA, LIA) readings.

- RGB-ITR electrical signals are heavily affected by noise.
- Make accurate measurement in a “disturbed” environment is not trivial: we will point out the all situations where the reading of the instruments cannot be used directly.

Fälschungsmöglichkeit

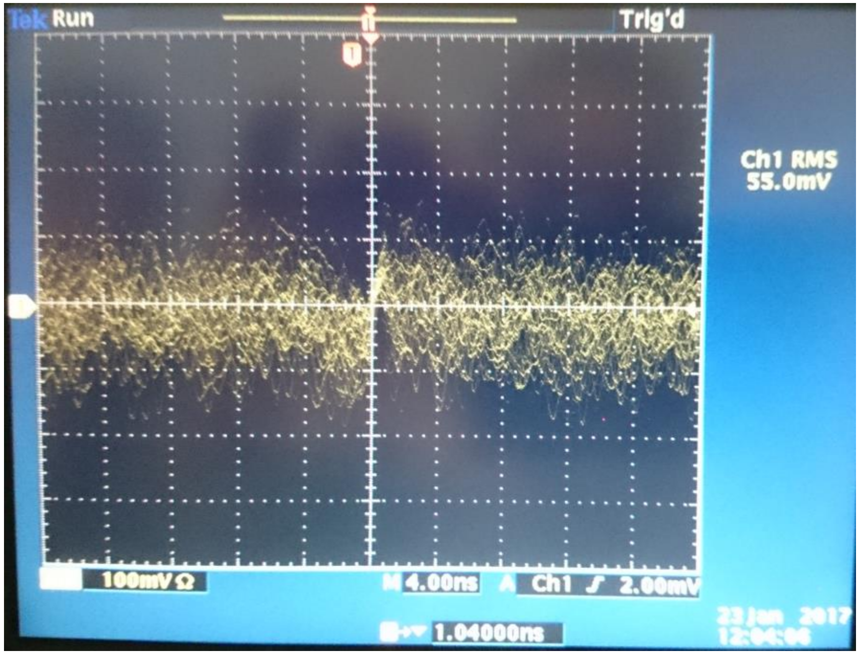
- Observe a **phenomena**
- Make some **hypothesis**
- Build a (mathematical) **model**
- Design a **refutable** experiment
- Make some **measurements**
- Check measurements **vs** hypothesis and model
- Draw the conclusions

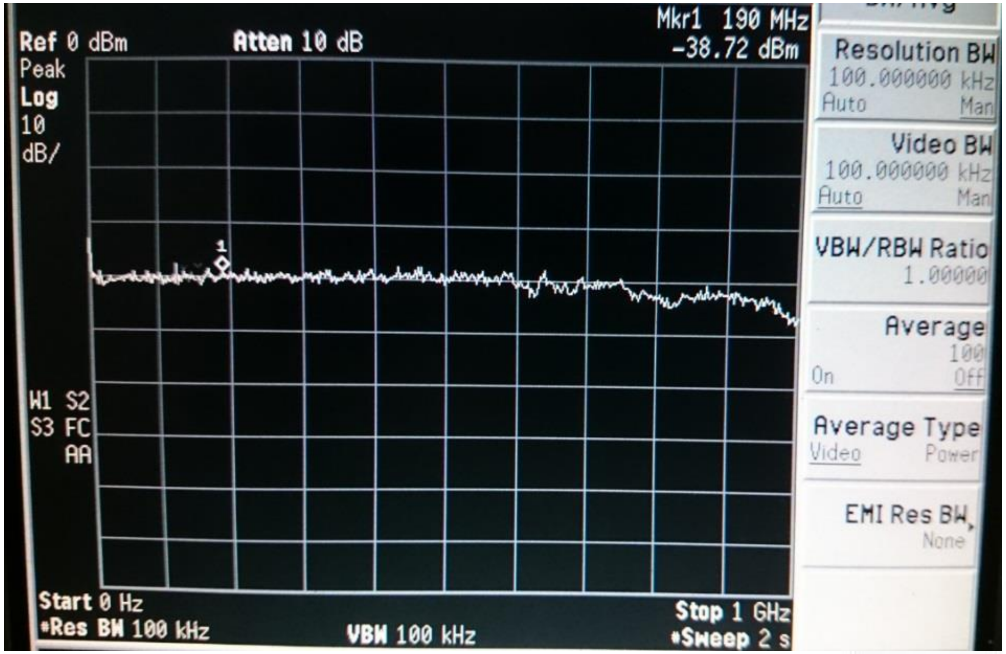
- Science is based on **errors**: the perimeter of the truth is defined by the errors.
- Progress is not the accumulations of (axiomatic) certainties.
- Progress is the systematic reduction and/or elimination of the errors.
- Biological evolution and closed loop control systems work in that way.



RGB-ITR electrical signals (RED channel)

Received signal Sr: Where is the signal?





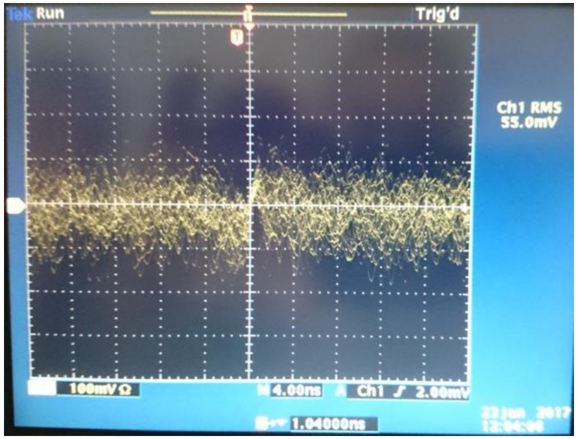
Modulating signal Fm: a perfect sinusoid at 190 MHz?

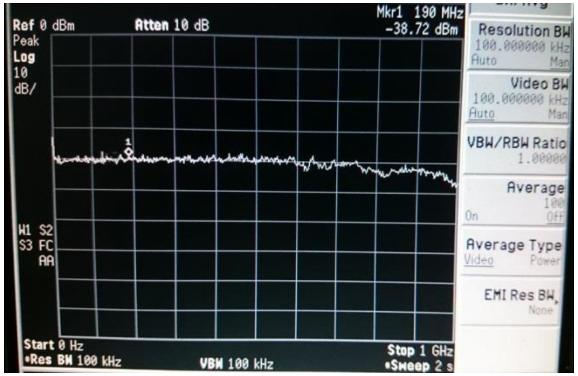
No, a very distorted square wave! A passive LPF fix this “issue”.

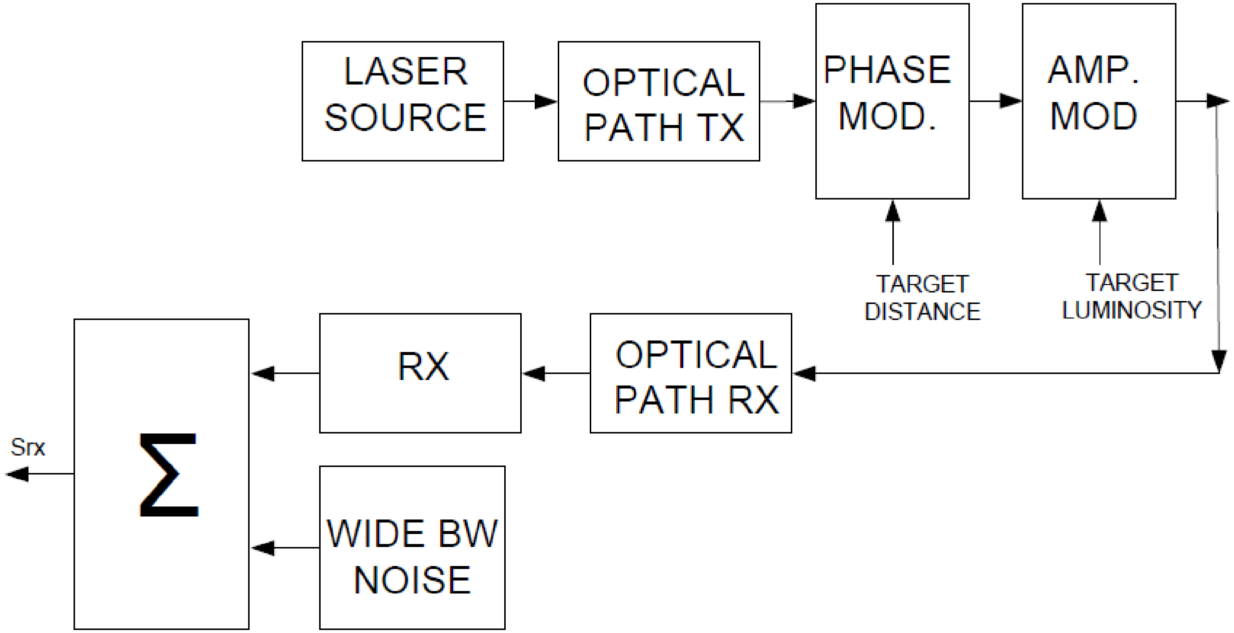
“It is not a bug, It’s a features”.

RGB-ITR electrical signals (RED channel)

Received signal Srx: Where is the signal?







RGB-ITR Measurement:: Scopes and Planning

- We need to measure (with the maximum possible precision/accuracy) **AMPLITUDE** and **PHASE** of a sinusoidal signal with a carrier frequency **F_m=190 MHz**
- AMPLITUDE and PHASE are modulated with a maximum bandwidth of **BW=3 kHz** (approx.)
- Received signal is heavily affected by noise: usually the received signal is dominated by noise.

Conclusions

- The measurements must be performed with radio frequency instruments and techniques.
- Special care must be dedicated to the proprieties of the noise component of the received signal.

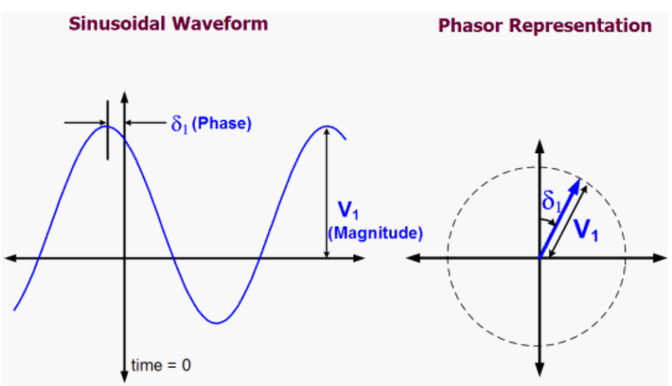
Received signal = AM/PM sinusoidal signal + thermal noise

Sinusoidal signal

Amplitude: measure the peak and divide by $\sqrt{2}$;

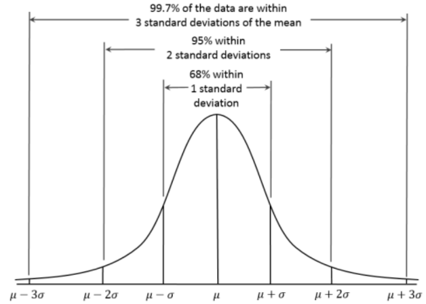
$$V_{RMS} = V_P/\sqrt{2}$$

Phase: set a reference point and measure Δt; $\varphi = 2\pi \frac{\Delta t}{T}$

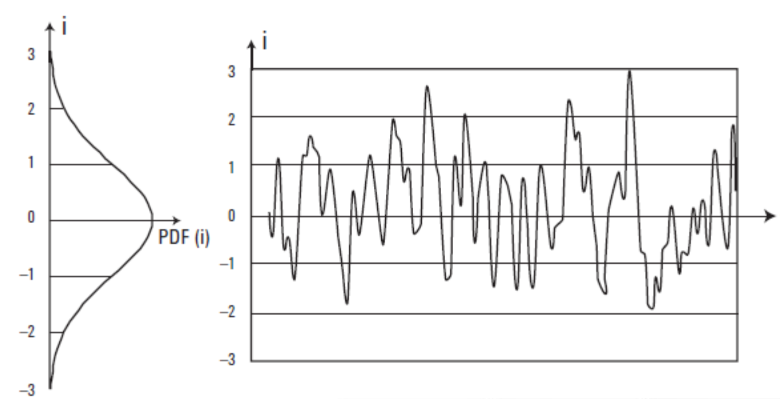


Noise signal

Amplitude: measure the peak (or p-p) and divide by an (arbitrary factor) 3-7.



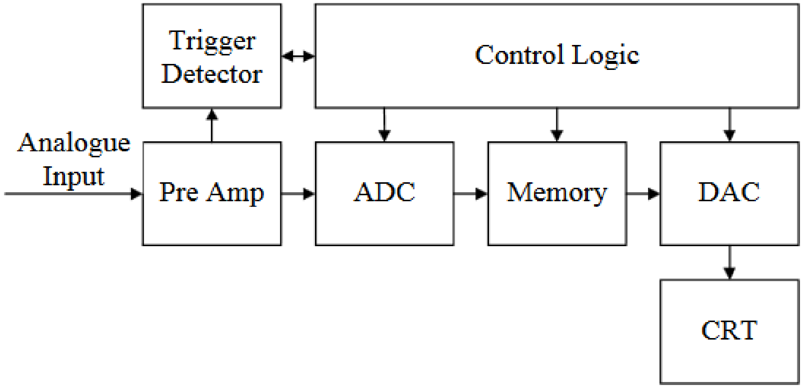
$$V_{RMS} = \sigma = \frac{V_P}{3} = \frac{V_{p-p}}{6} \quad \text{(approx)}$$



RGB-ITR Measurement:: Scopes and Planning

Time domain and frequency domain

The **OScilloscope** (OS) produces accurate measurements of sinusoidal signals or noise (with limited precision) but it is useless for very noisy signals.



The **Spectrum Analyzer** (SA) allows precise measurements of amplitude of sinusoidal signals in presence of noise and very accurate measurement of wideband noise. Unfortunately, standard SA discards the **PHASE** information and its usage is not straightforward as an OS: there are a lot of parameters to set and the measures are expressed in ***dBm*** on a fixed (adjustable) bandwidth.

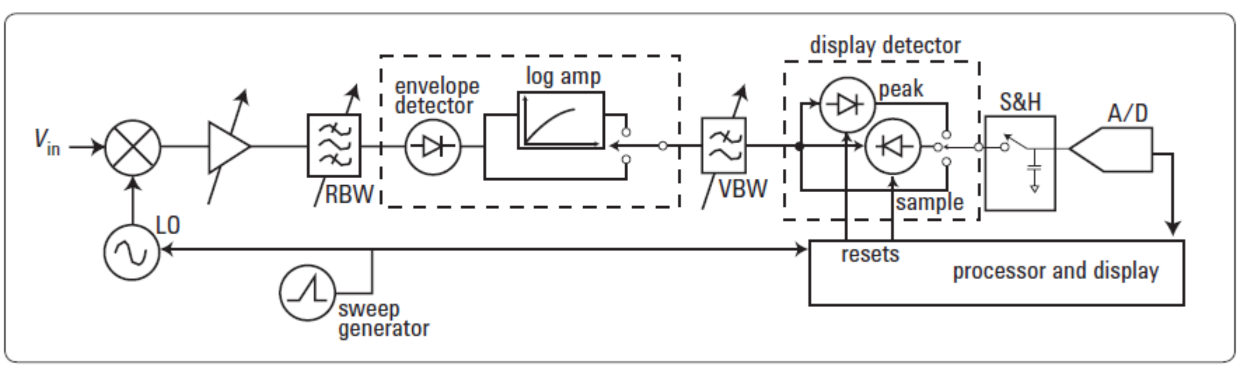
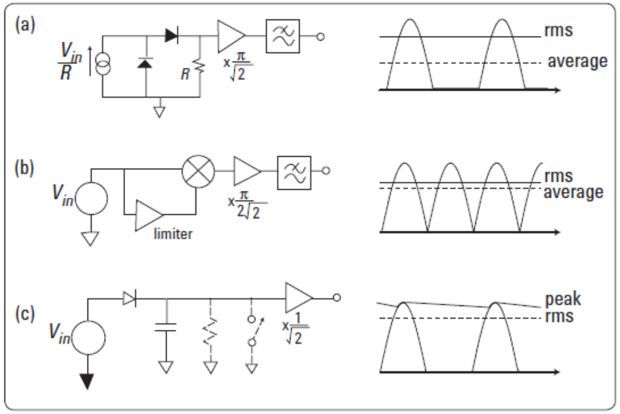


Figure A. Simplified spectrum analyzer block diagram



Radio Frequency Measurement Basics

- Interconnection (cables) are not transparent: they may modify amplitude and phase of the signals.
- Connections & Connectors: there are _*very*_ critical (wearing).
- All signals must be terminated or doubly terminated (Z_0) because *signal reflections degrade the accuracy*.
- Signals and circuits must be protected from external and internal disturbances: no shielding effort is too paranoid.
- Power supply lines must be electromagnetically decoupled because they may become *undesired inputs for external unwanted signals*.

Conclusions

- Instruments and accessories used for the measurements take active part of the measure: you cannot just plug it and read the display. ***Measurement planning is not an option***.
- ***Electromagnetic compatibility (EMC) is not an option***.

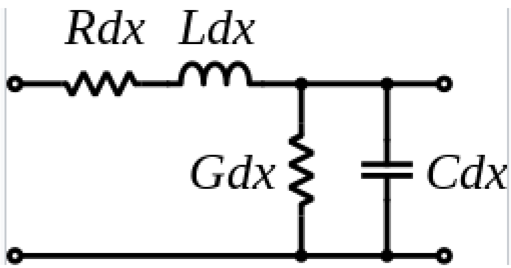
Radio Frequency Measurement Basics

Everything is a transmission line

As first design step, consider the transmission line lossless: R=0, G=0. In this case, the characteristic impedance is.

Transmission lines move **INFORMATION** as ELECTRICAL **POWER** (voltage, current) from one place to another.

$$Z_o = R_o = \sqrt{\frac{L}{C}}$$



$$Z_0 = \frac{138}{\sqrt{k}} \log \frac{d_1}{d_2}$$

Where,

Z_0 = Characteristic impedance of line

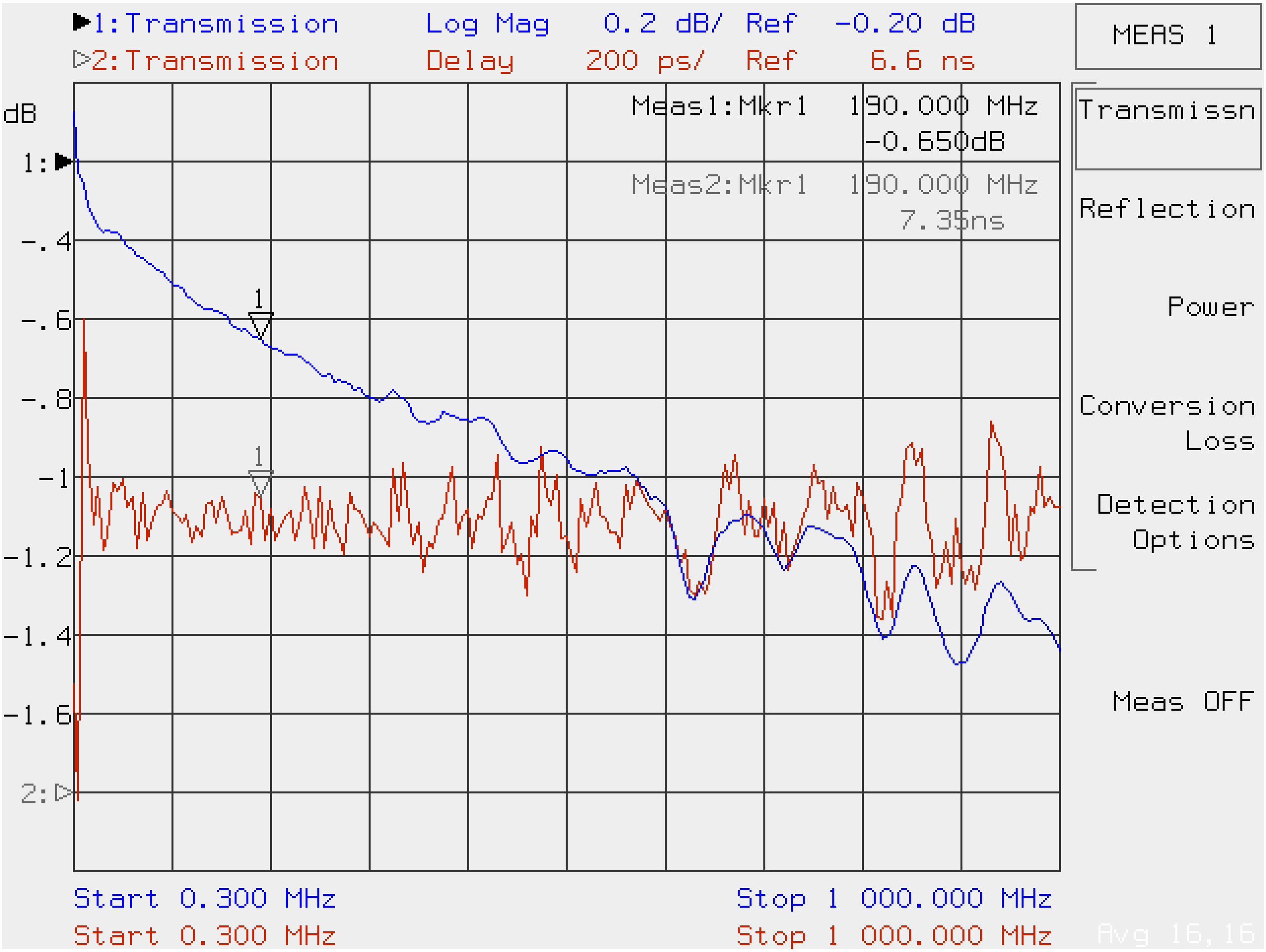
d_1 = Inside diameter of outer conductor

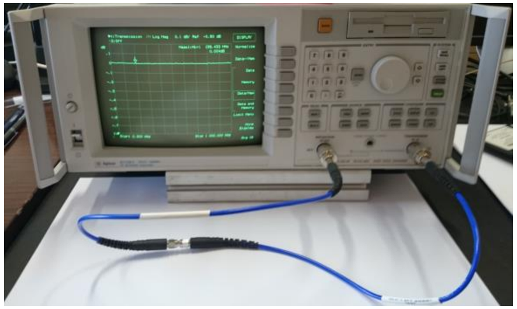
d_2 = Outside diameter of inner conductor

k = Relative permittivity of insulation between conductors

Everything is a transmission line: (HP/A 8712ET VNA)

50 Ohm RG-58 cable with SMA connectors, L=110 cm





Radio Frequency Measurement Basics: Signal Power

If a lossless transmission line is connected to a matched resistive load, the received power is

$$P_o = \frac{V^2}{R_o} = I^2 R_o$$

In radio frequency measurement, everything is calibrated in ***dBm*** (decibel referred to P_r=1mW over Ro=50 Ohms. ***Why?***

$$P_o[dBm] = 10\ log\left(\frac{P_o}{P_r}\right) = 10\ log\left(\frac{\frac{(V_o)^2}{R_o}}{\frac{(V_r)^2}{R_o}}\right) = 20\ log\left(\frac{V}{Vr}\right)$$

Radio Frequency Measurement Basics: Noise Power

If a lossless transmission line starts from a matched source (Zo=Ro), the output impedance at the load is Ro, therefore the equivalent thermal noise power of the source is

$$P_N = K_B\, T\, NBW \qquad \text{[Watt]}$$

K_B: Boltzmann’s constant = 1.38E-23 [Joules/Kelvin]

T : absolute temperature = 300 [K] (standard lab. cond.)

NBW: equivalent noise bandwidth [Hz , 1/sec]

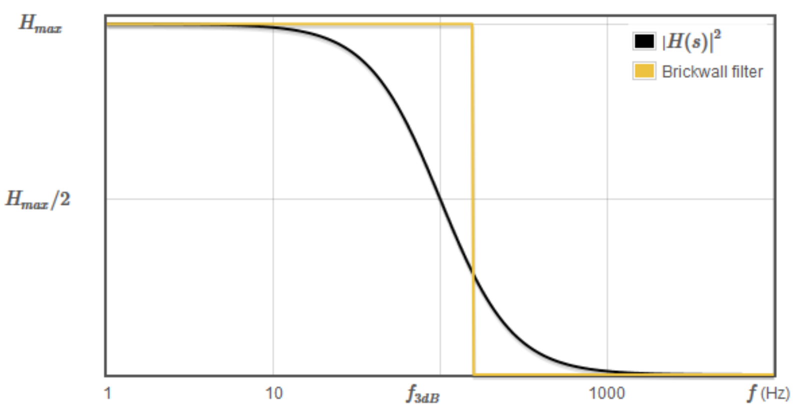
P_N: total noise power (Watt) over the bandwidth NBW (Hz)

Another key parameter is the noise power density “ p_n”

$$p_n = \frac{P_N}{NBw} = K_B\, T \qquad \text{[Watt/Hz]}$$

Radio Frequency Measurement Basics: NBW

Every physical system has a low-pass (or band-pass) frequency response (transfer function). Given a system with a transfer function H(s), the spectrum of any input noise will be shaped by it. Under white noise (with equal “pn” noise density over all frequencies), it is convenient to replace the transfer function by a "brickwall" filter, in which the noise power is the same up to a certain frequency NBW (Δf) and after that is zero. That frequency is defined so that the total output noise power is the same for both the system and the brickwall filter. Thus, the area under this rectangle must be the same as the area under the original system.



$$A_{H(s)} = A_{rectangle}$$

Order	Δf
1	1.57 f_{3dB}
2	1.22 f_{3dB}
3	1.15 f_{3dB}
4	1.13 f_{3dB}
5	1.11 f_{3dB}

$$P_N = K_B\, T\, NBW$$

Radio Frequency Measurement Basics:: [dBm]!

dBm	Watt	V(RMS)	V_p
+30	1 Watt	7.07 V	10.0 V
+20	100 mW	2.24 V	3.16 V
+10	10 mW	707 mV	1.00 V
0	1 mW	224 mV	316 mV
-10	100 μW	70.7 mV	100 mV
-20	10 μW	22.4 mV	31.6 mV
-30	1 μW	7.07 mV	10.0 mV
-40	100 nW	2.24 mV	3.16 mV
-50	10 nW	707 μV	1.00 mV
-60	1 nW	224 μV	316 μV
-70	100 pW	70.7 μV	100 μV
-80	10 pW	22.4 μV	31.6 μV
-90	1 pW	7.07 μV	10.0 μV
-100	100 fW	2.24 μV	3.16 μV
-120	10 fW	707 nV	1.0 μV
-150	1 fW	224 nV	316 nV
-160	100 aW (atto)	70.7 nV	100 nV
-180	10 aW	22.4 nV	31.6 nV
-190	1 aW	7.07 nV	10 nV
-200	100 zW (zepto W)	2.07 nV	3.6 nV

The dBm scale was proposed in 1940 for telephony (Ro=600 Ohm) and it has been adopted for radio frequency measurements (Ro=50 Ohm).

<u>Why 50 Ohm</u>?

Experimentation in the early 20th century determined that the best power handling capability could be achieved by using 30 Ohm coaxial cable, whereas the lowest signal attenuation could be achieved by using 77 Ohm coaxial cable. However, there are few dielectric materials suitable for use in a coaxial cable to support 30 Ohm impedance. Thus, 50 Ohm characteristic impedance was selected as the ideal **compromise**, offering good power handling and low attenuation characteristics.

75 Ohm cables are used for television signals (low attenuation). Low impedance cables (30 Ohm or less) are used for high power applications.

RGB-ITR RED channel receiving channel analysis

- Oscilloscopes (time domain instruments) are useless when the SNR is too low. You need to switch to spectrum analyzers (SA).
- SA works in the frequency domain using incoherent signal demodulation: the phase information is LOST.
- SA uses a sweep local oscillator, up/down conversion mixers and programmable narrow band filters: a lot of parameters have to be set in order to make a good (amplitude only) measurement.
- SA may have very high frequency coverage (1Ghz), good selectivity (100 kHz) and very good dynamic range (80 dB) but all the parameters are interlinked: make a mistake is very easy.
- Last but not least, if you connect an unknown signal to the SA input, there is a serious probability that you damage permanently the instrument: SA input circuits (mixer) is not protected from overload! (1 Watt is enough to damage permanently the input).

RGB-ITR RED channel::SA common settings

After few iteration we choose the following SA measurement settings (fixed over all the measurement taken)

- Frequency Span: 0-1GHz (real F_{min}=9 kHz)
- Bandwidth: 100 kHz (video BW=resolution BW = noise BW)
- Sweep time: 1-2s
- Input Attenuation: 10 dB (for protection and dynamic range matching)

Nota Bene

- Increase the frequency span is not interesting because the APD detector has 1GHz BW and the following LNA has 0.5 GHz BW.
- SA bandwidth filter cannot be reduced further (less than 100 kHz) with reasonable scan time.

RGB-ITR RED channel:: SA specific settings

The SA internal demodulator must be set differently if in case of sinusoidal signals or noise:

Peak detection is used primarily when measuring sinusoidal (spectral) components. Peak detection obtains the maximum video signal value between the last display point and the present display point and stores this value in memory. This detection should not be used for noise measurement because the peak detector is calibrated ($V_{RMS} = V_P/\sqrt{2}$) for sinusoidal signals. For noise-like signal $V_{RMS} = V_P/3$ (approx.). Peak detection is selected at power on BY DEFAULT.

Sample detection is used primarily to display noise or noise-like signals. *This detection should **not be used** to make the most accurate amplitude measurement of non noise-like (e.g. sinusoidal) signals*. In sample mode, the instantaneous signal value at the present display point is placed in memory. Usually, the "raw" measurement is too noisy: some averaging (10-100, typ. 30) factor must be set using the average "Power" setting.

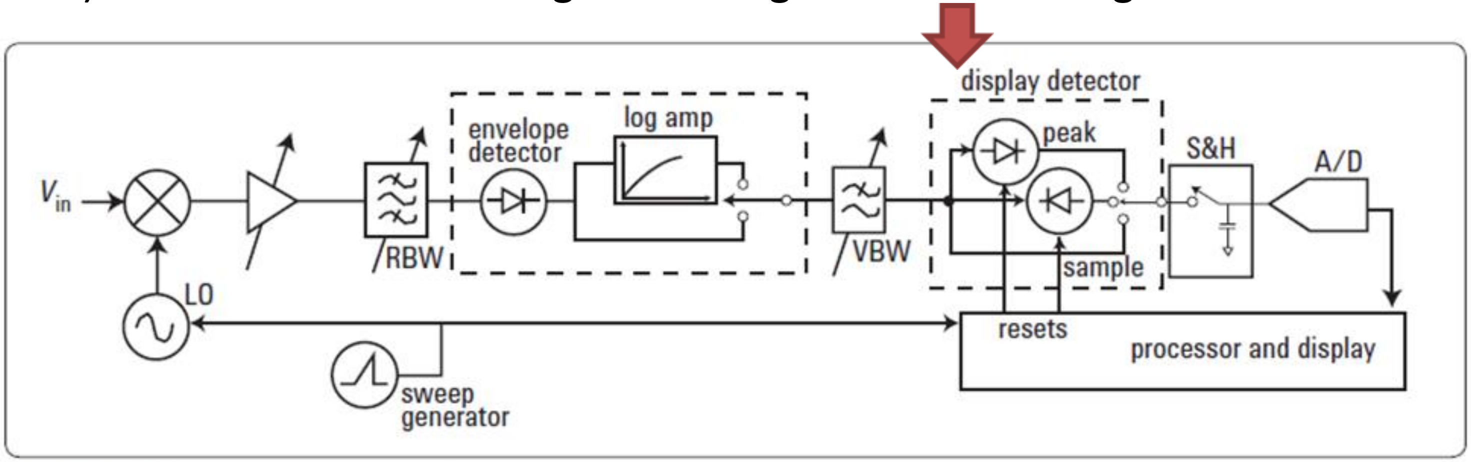
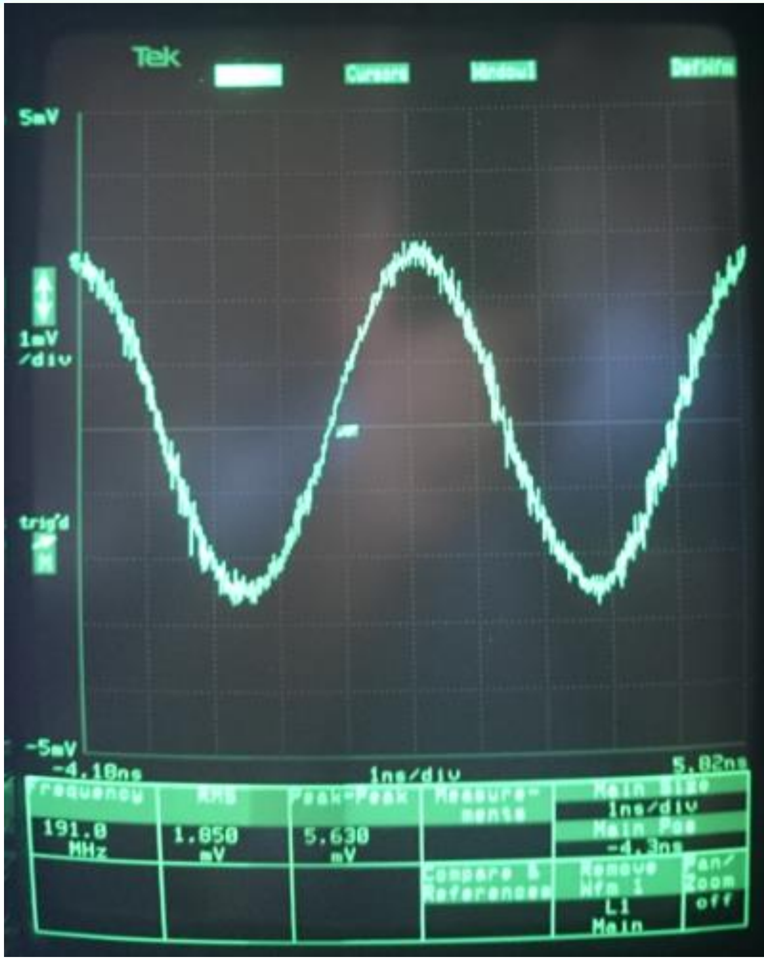
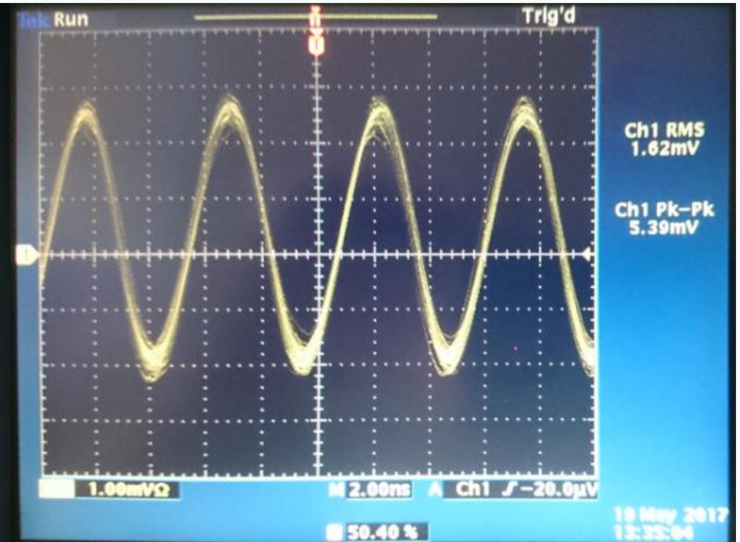


Figure A. Simplified spectrum analyzer block diagram

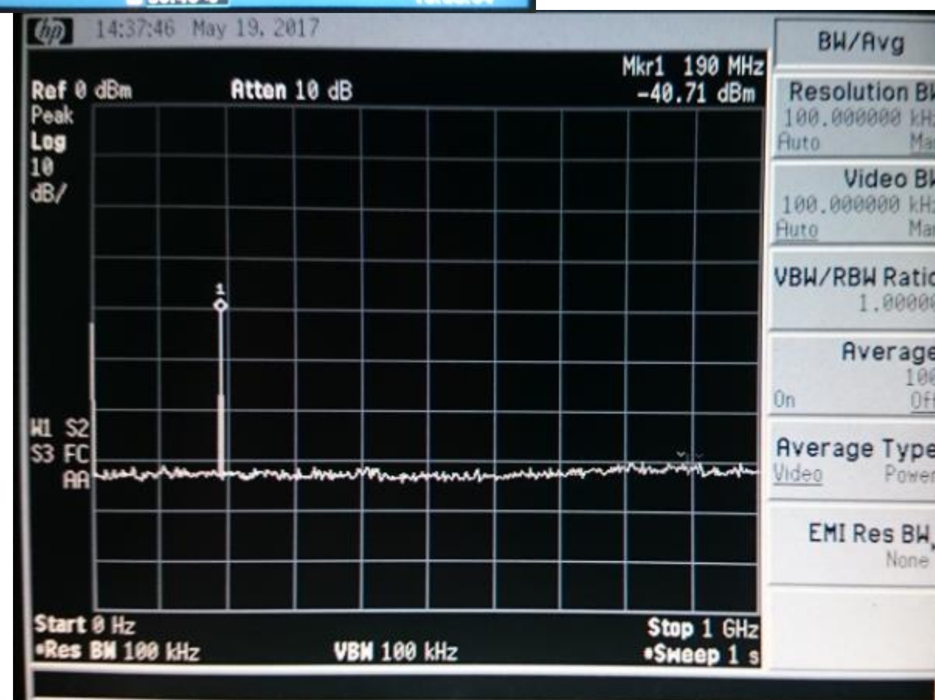
Sinusoidal signal:: Fm=190 MHz, A=-40dBm





TEK TDS3034

300MHz, 8 bit

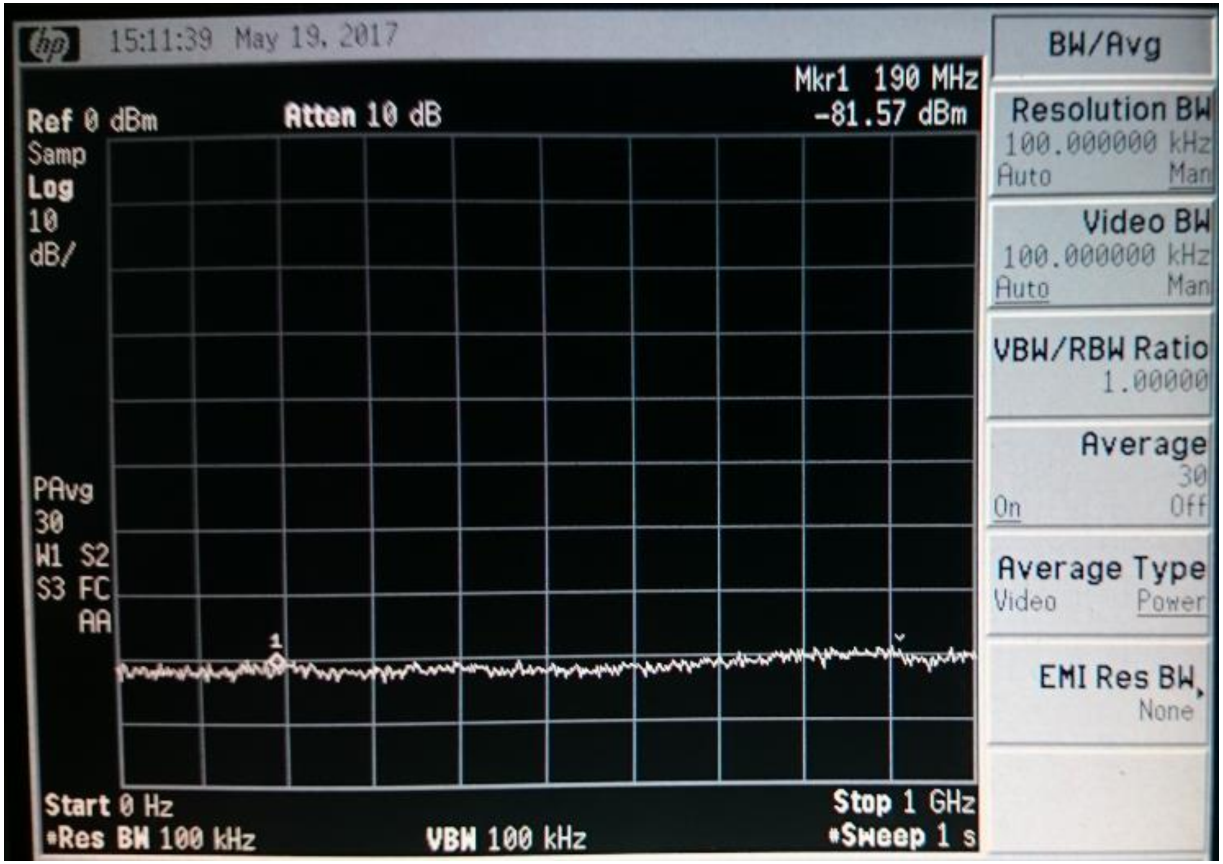


TEK 11402 - 600MHz, 10 bit

HP SA E4411B – 1.5GHz

BW 100 kHz, 90 dB

Spectrum Analyzer Noise Floor



NBW = 100kHz

Video BW=100 KHz

Average=30 (“Power”)

P_{nSA}= -80 dBm = 10 pW

$p_{nSA} = \frac{P_{nSA}}{NBW}$ = 0.1 fW/Hz

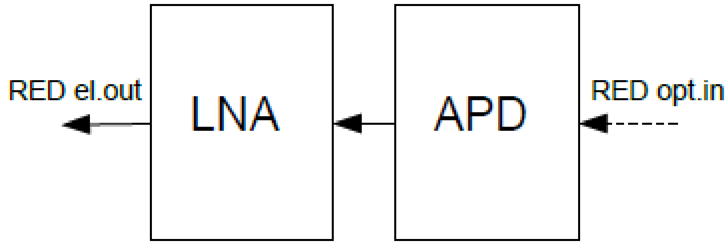
$p_{nTH} = K_B T$ = 4.14E-21 W/Hz

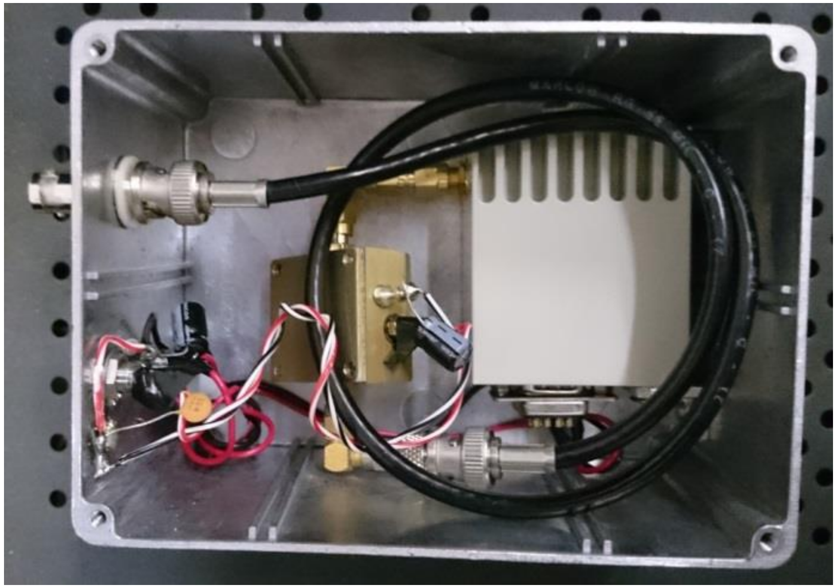
NF(SA) = $\frac{p_{nSA}}{p_{nTH}}$ = +43.8 dB!

Despite its internal noise (noise figure = 44 dB), a spectrum analyzer has better performances than an oscilloscope measuring the <u>amplitude</u> of high dynamic range signals in a very noisy environment.

RGB-ITR RED channel architecture

The RGB-ITR RED channel receiver is composed by an HAMAMATZU C5658 “APD” module and a Mini Circuits ZFL-500LN+ “LNA” low noise amplifier in cascade.





MODULE

APD module C5658

Detects optical signals at 1 GHz, with high sensitivity

APD module C5658 is a highly sensitive photodetector consisting of a Si APD (avalanche photodiode), a bias power supply and a low-noise amplifier, all integrated into a compact case. The APD used has an effective active area of ϕ0.5 mm to allow efficient coupling to a light beam in applications such as spatial light transmission. The APD internally multiplies the photocurrent to produce an ample gain (set to 100 times for C5658) and also features high-speed response, achieving detection limits up to 1 GHz wideband and -48 dBm (16 nWr.m.s.) noise level in combination with the low-noise amplifier. C5658 also incorporates a thermosensor and a temperature-compensated bias power supply necessary for stable operation of the APD. Highly sensitive optical measurements can be made just by supplying +12 V to C5658.

Features
- High-speed response and high sensitivity
- Flat frequency characteristics
- Compact and lightweight
- Single power supply operation

Applications
- Laser radar
- Spatial light transmission
- Optical rangefinder

Low Noise Amplifier ZFL-500LN+

50Ω 0.1 to 500 MHz

SMA version shown
CASE STYLE: Y460

Connectors	Model
SMA	ZFL-500LN+
BNC	ZFL-500LN-BNC+
BRACKET (OPTION “B”)	

+RoHS Compliant
The +Suffix identifies RoHS Compliance. See our web site for RoHS Compliance methodologies and qualifications

Features
- very low noise, 2.9 dB typ.
- good VSWR, 1.5 :1 typ.
- protected by US Patent, 6,943,629

Applications
- VHF/UHF
- small signal amplifier
- communications system

Low Noise Amplifier Electrical Specifications

MODEL NO.	FREQUENCY (MHz)		NOISE FIGURE (dB)	GAIN (dB)		MAXIMUM POWER (dBm)		INTERCEPT POINT (dBm)	VSWR (:1) Typ.		DC POWER	
	f_L	f_U	Typ.	Min.	Flatness Max. Total Range	Output (1 dB Compr.)	Input (no damage)	IP3 Typ.	In	Out	Volt (V) Nom.	Current (mA) Max.
ZFL-500LN+	0.1	500	2.9	24	±0.5	+5	+5	+14	1.5*	1.6	15	60

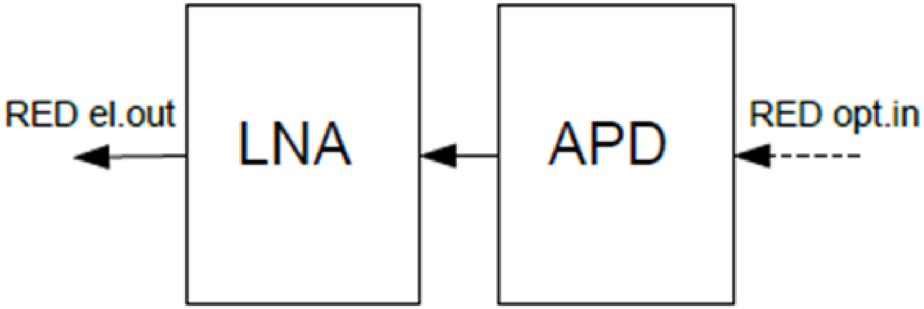
m = mid range [2 fL to fU/2]

RGB-ITR RED channel:: NOISE estimation from specs

The APD module has a total noise output (over a 1 GHz BW) of $P_N =$ -48 dBm. The noise figure of the LNA is 3 dB (the double) higher of the equivalent thermal noise of a 50 Ohm resistor; the equivalent LNA input noise is:

$$P_N = 2\, K_B\, T\, NBW = 13 \text{ pW} = -79 \text{ dBm}$$

We consider the internal noise of the LNA negligible respect the output noise of the APD.



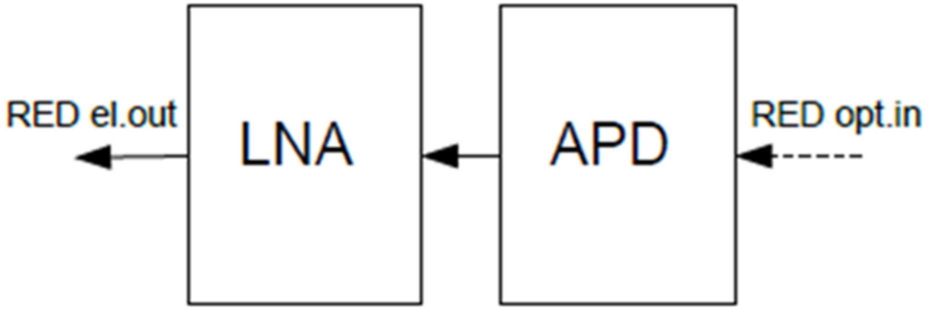
RGB-ITR RED channel:: NOISE estimation from specs

The noise power at the LNA output is the noise power input (from the APD) multiplied by the LNA gain (+24 dB); the output power is:

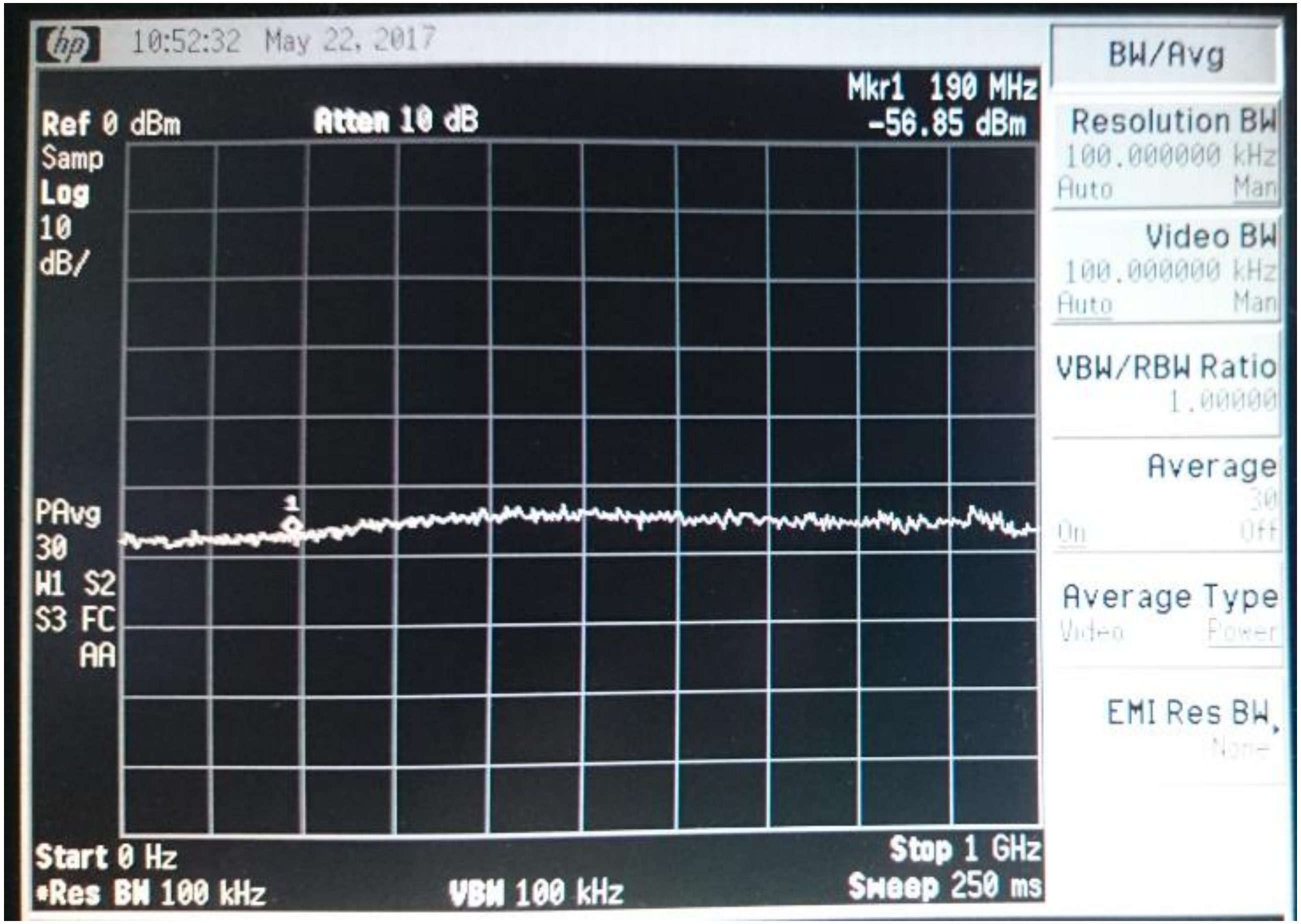
P_{N-LNA_out} = -48 + 24 = -24 dBm = 3.98 μW=14.1 mV(RMS)

The estimated noise power density at LNA output over 1.0GHz (-3dB) bandwidth is:

$$p_{n-LNA} = \frac{P_{N-LNA}}{NBW} = \mathbf{2.53\ fW/Hz}$$



RGB-ITR RED channel receiving channel: NOISE



NBW = 100kHz

Video BW=100 KHz

Average=30 (Power)

P_{nSA}= -57 dBm = 2 nW

$p_{nSA} = \frac{P_{nSA}}{NBW}$ = **20 fW/Hz**

The noise power spectral density is basically constant over 1GHz bandwidth but its measured value is (approx.) **TEN** times the value computed from data sheet (**2.5 fW/Hz**).

We have verified that the LNA has the declared (negligible) noise figure.

RGB-ITR RED channel receiving channel: NOISE

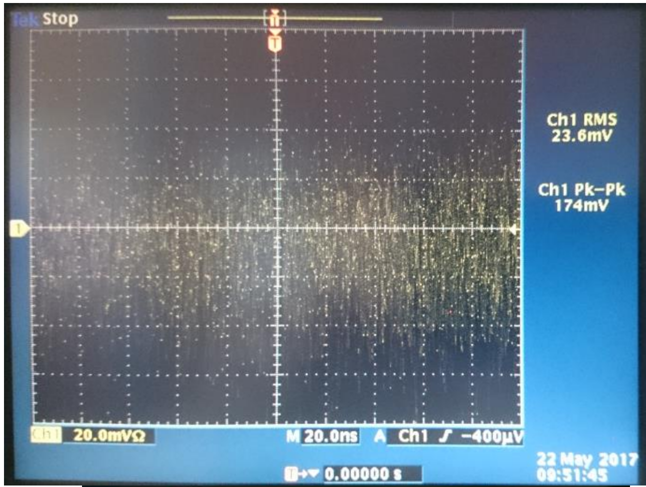
ENEA AGENZIA NAZIONALE PER LE NUOVE TECNOLOGIE, L'ENERGIA E LO SVILUPPO ECONOMICO SOSTENIBILE

The SA noise measurement have been duplicated using two wideband scopes:

TEK 3034, BW = 300MHz

$V_{RMS} = 23.4;\ V_{pp} = 174mV; pf = V_{pp}/V_{RMS}$=7.57

$P_N = 11.1\ \mu W; p_n =$ **23.7 fW/Hz**

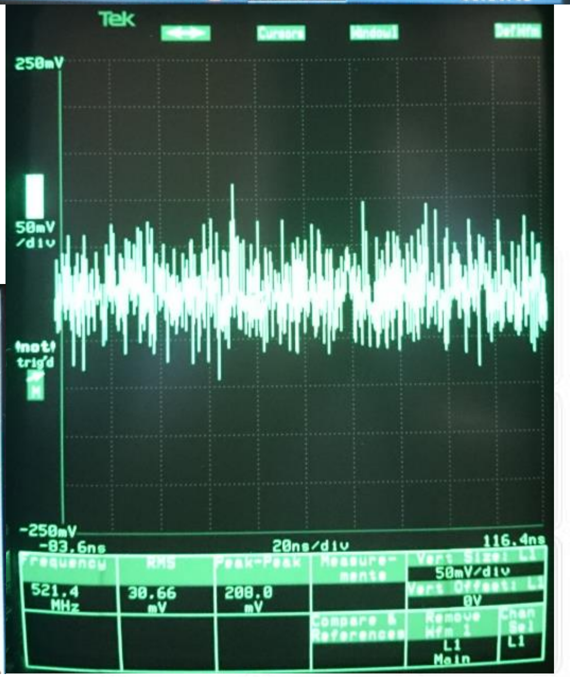


TEK 11401, BW = 600MHz

$V_{RMS} = 30.66;\ V_{pp} = 208.0\ mV; pf = V_{pp}/V_{RMS} = 6.78;$

$P_N = 18.8\ \mu W; p_n =$ **20.0 fW/Hz**

Oscilloscope measurements confirm spectrum analyzer results with very good agreement.

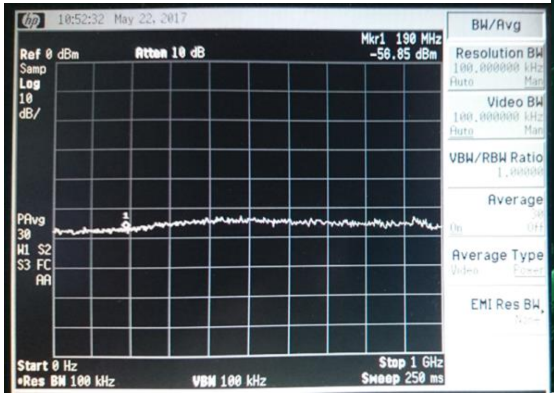


The real motives behind the discrepancies between the expected noise power and the real measured values are still **unknown** and should be investigated.

HP E441B, BW=100kHz

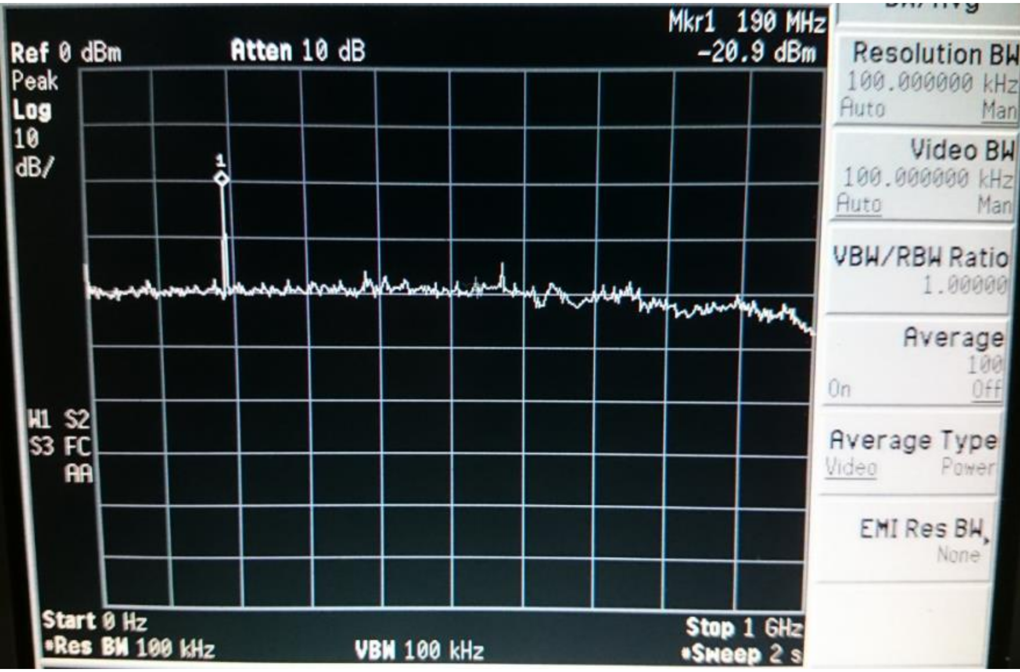
$P_N = -57\ dBm = 2.0\ nW;$

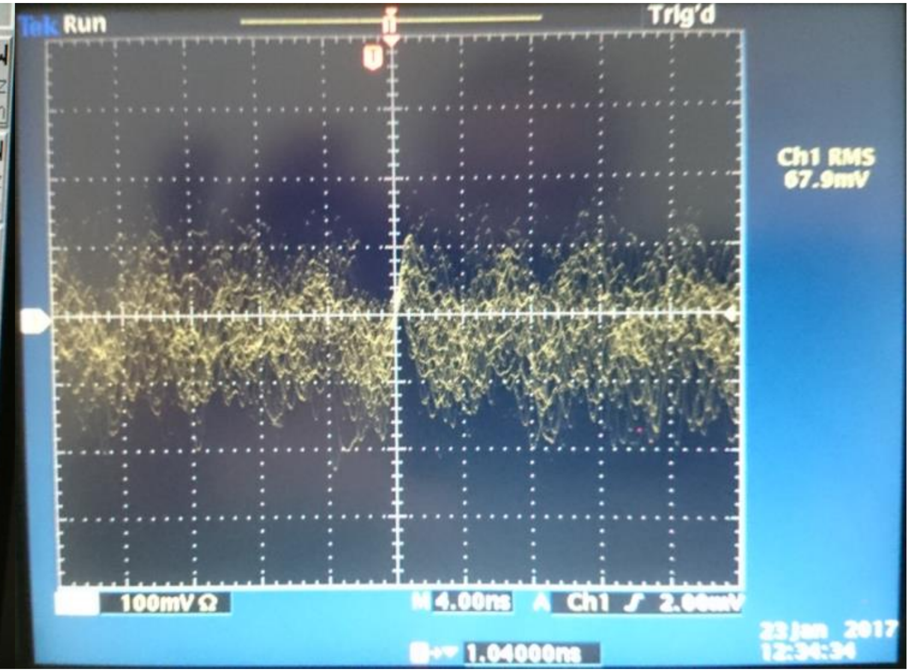
$p_n =$ **20.0 fW/Hz**



RGB-ITR RED channel receiving channel analysis

Typical situations using RGB-ITR





Signal power = -20.9 dBm = 20.1mV = 8.13 µW

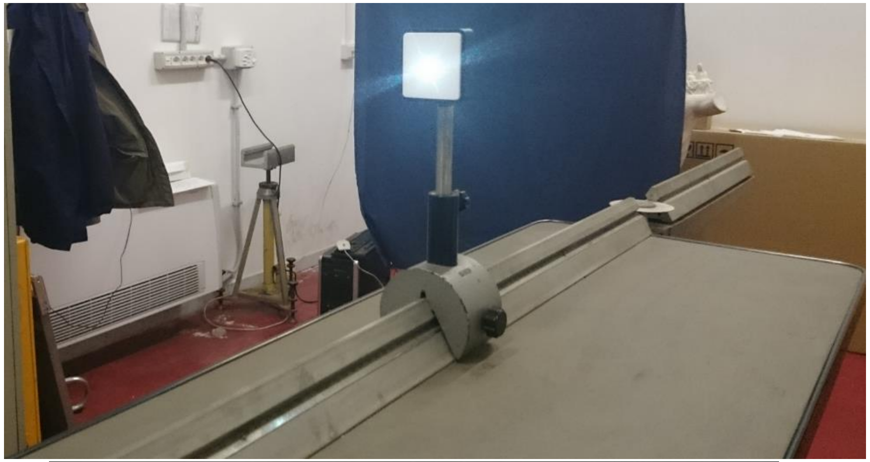
Noise power = 23.6 mV = 11.1µW

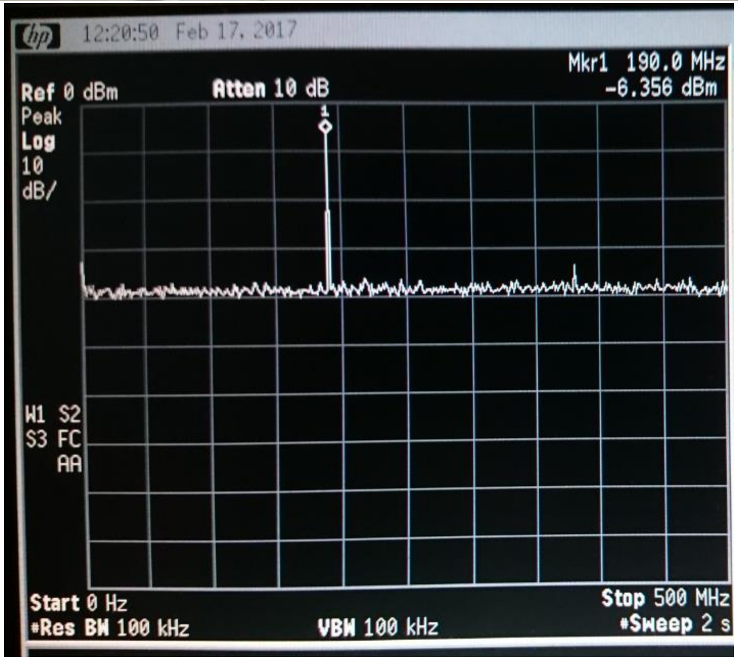
Measured signal+noise power using the OS: 92.2 µW

Wrong measure? *(This measure must be done again)*

RGB-ITR RED channel receiving channel analysis

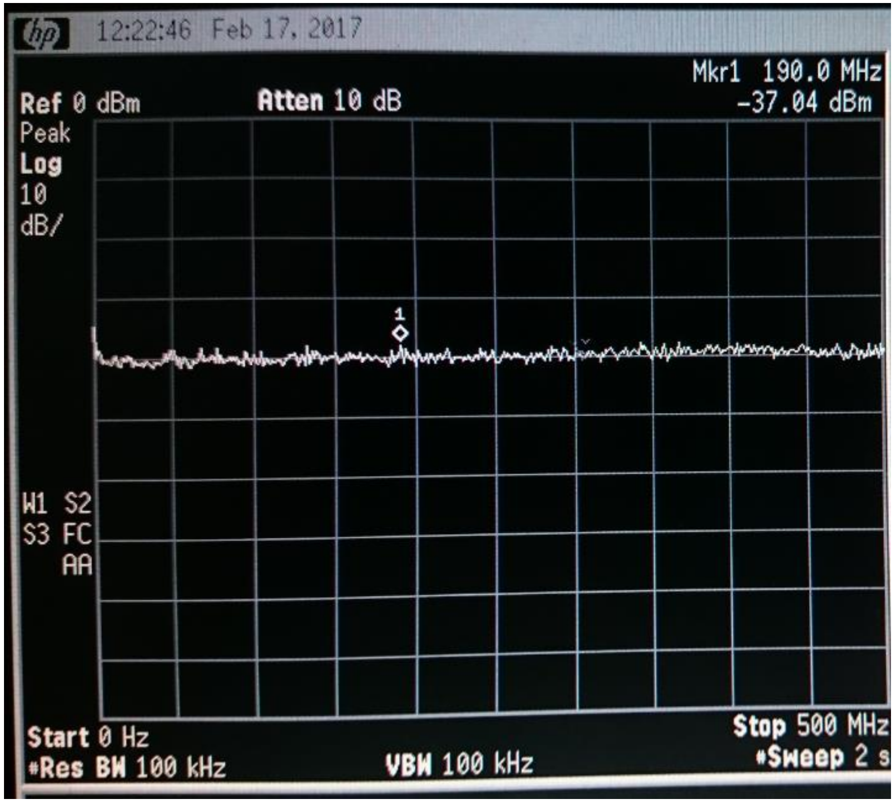
RED channel static dynamic range test bed: WHITE target





RGB-ITR RED channel receiving channel analysis

RED channel static dynamic range test bed: BLACK target



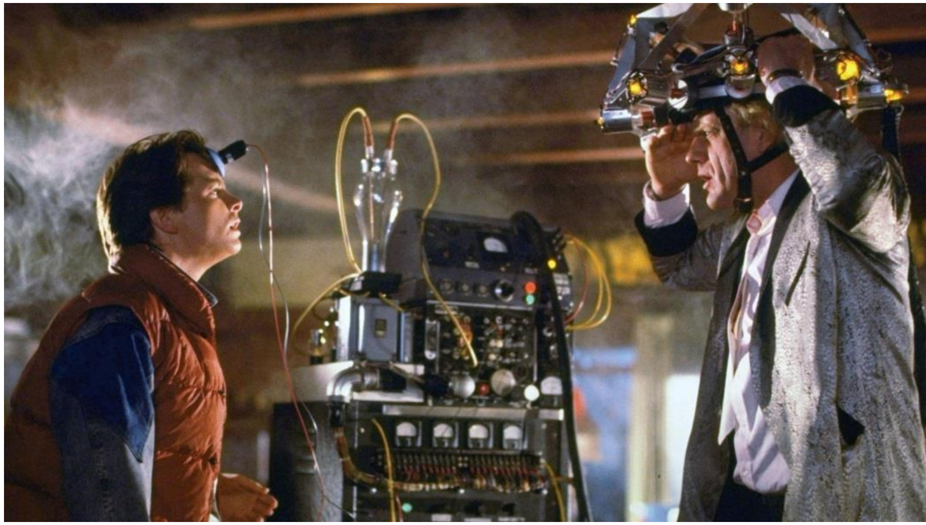
TARGET color:

WHITE = - 6 dBm ; BLACK = - 37 dBm

Target color variation dynamic range = 30-40 dB

RGB-ITR RED channel receiver::SUMMARY

ENEA AGENZIA NAZIONALE PER LE NUOVE TECNOLOGIE, L'ENERGIA E LO SVILUPPO ECONOMICO SOSTENIBILE



- Wavelength : 660 nm (visible RED)
- Modulation frequency Fm: 190 MHz
- Folding distance L = 0.79 m
- Bandwidth: less than 3 kHz
  (max LIA BW =1600 Hz, tau=100 us)
- Minimum received signal: 1 mV
- Maximum received signal: 100 mV
- Typical useful dynamic range = 40 dB
- Background noise: white noise, flat up to 1 GHz, $p_n = 20\ fW/Hz$

Oscilloscope: **useless**; Spectrum Analyzer: **useless** (phase information is lost).

OBSERVATION: the modulating (carrier) frequency is F_M=190 MHz but the really useful signal bandwidth is 3 kHz!

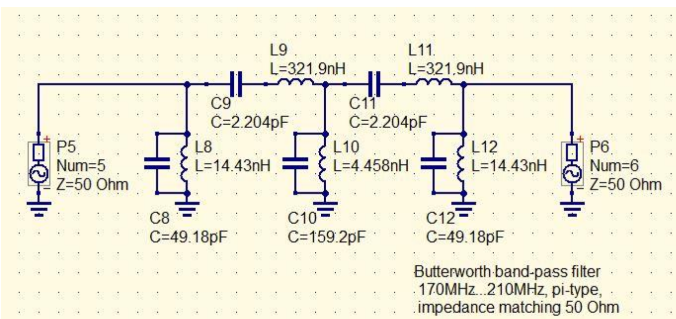
IDEA!: insert a narrow band pass filter around 190 MHz with a +/-10 (20 kHz BW)

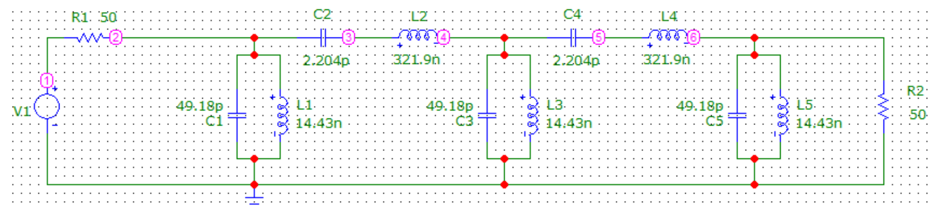
RED channel receiver upgrade:: narrow band pass filter

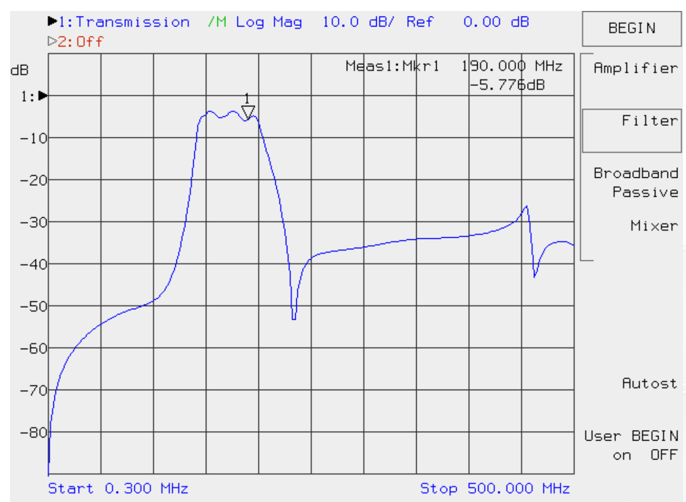
- $F_o = 190\ MHz$
- $BW = 20\ kHz$
- $Q = \frac{F_o}{BW} = 9500\ !$

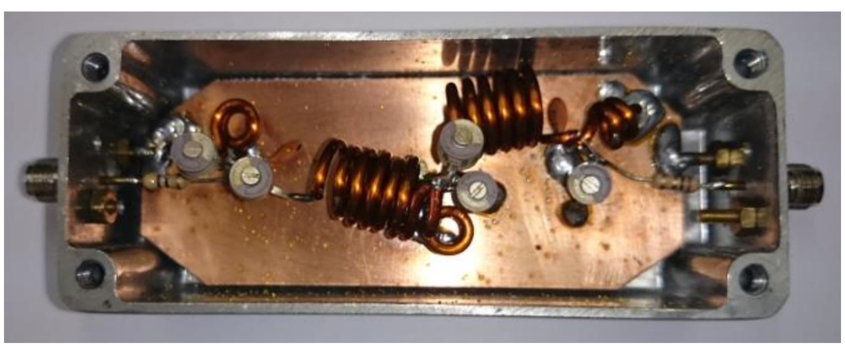
Q = 10.000 is an impossible target to achieve with analog passive filter because the inevitable parasitic parameters of the REAL components.

Design a narrow band pass filter at 190 MHz needs the right combination of science and art (“The Art of Analog Design”): CAD, simulations and a lot of practice.









RGB-ITR receiver and signal processing

Narrow band pass filter fails, but the idea is valid.

How to measure amplitude and phase of a narrow band signal covered by wideband noise?

IDEA: periodic, narrowband signal? **Fourier Transform**!

Jean Baptiste Joseph **Fourier** est un mathématicien et physicien français né le 21 mars 1768 à Auxerre et mort le 16 mai 1830 à Paris. Il est connu pour avoir déterminé, par le calcul, la propagation de la chaleur en utilisant la décomposition d'une fonction quelconque en une série trigonométrique convergente. De telles fonctions sont appelées séries de Fourier ; la méthode de calcul permettant de façon réversible de passer d'une fonction à la série trigonométrique correspondante est la **transformation de Fourier**. Cette méthode - très féconde - est devenue incontournable en théorie du signal, imagerie numérique, compression de données, dans l'exploitation des systèmes sans fils et ***pour réaliser des filtres (analogique et numériques) très sélectifs***.

RGB-ITR channel simulator

Why design and built a simulator? For four good reasons:

- RGB-ITR apparatus is used regularly for field scans, therefore the (low) availability is a critical factor;
- planning and executing direct measurements on RGB-ITR is complex, time consuming and the reproducibility is critical;
- Physical simulators, despite their limitations, are more easy to use than the real apparatus;
- Physical simulators cost less than the real apparatus.

Conclusions

- Data recovered from the previous experiments shown that reproducing the received signals is relatively easy.
- Build a channel simulator can accelerate the development of the lock-in prototype decoupling tests from the physical availably of RGB-ITR complete system.

RGB-ITR channel simulator

Specification of the simulator (for the RED channel):

- Static attenuation of signal: 0-60 dB (or better)
- Fast (at least 10 kHz bandwidth) variable attenuation 0-40 dB (voltage controllable)
- Fast (at least 10 kHz bandwidth) 0-360° variable phase delay (voltage controllable).
- Internal noise source (white noise up to 1 GHz)
- Static attenuation of noise: 0-60 dB or better.

RGB-ITR channel simulator block diagram

Phase Mod.: 0-360°, voltage controlled

Amp. Mod.: 0/-40 dB, voltage controlled

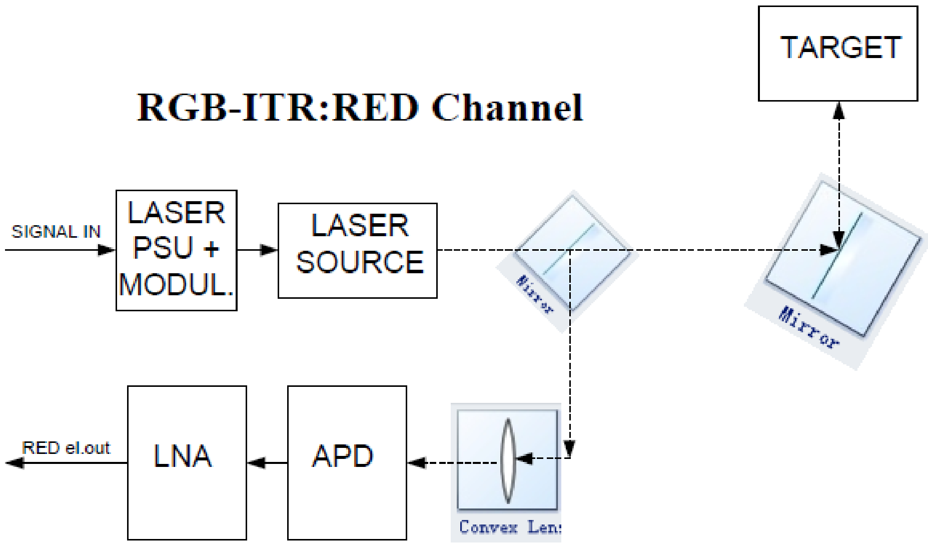
LNA: low noise amplifier (+14 dB); 10dB attn. + ZFL500 (+24 dB) amplifier (MC)

ATTN: TRILITC 0/110 dB wideband attenuator, 1 dB step, 1dB accuracy, 1GHz

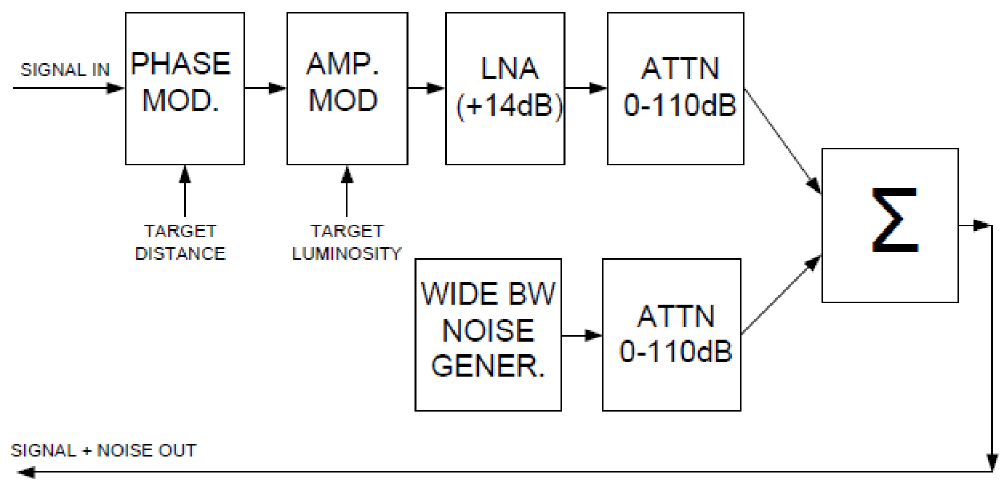
Wide BW Noise Gener.: ready-to-use, diode based noise generator.

Σ: ZSC 2-1+ power combiner (Mini Circuits)

RGB-ITR:RED Channel

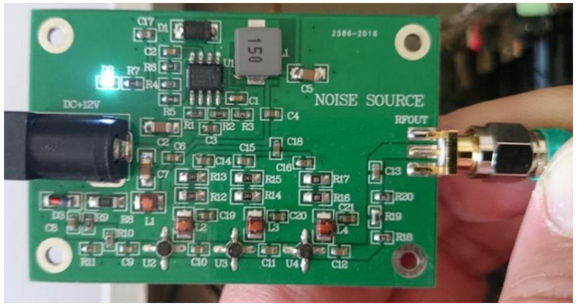


RGB-ITR:Channel Simulator

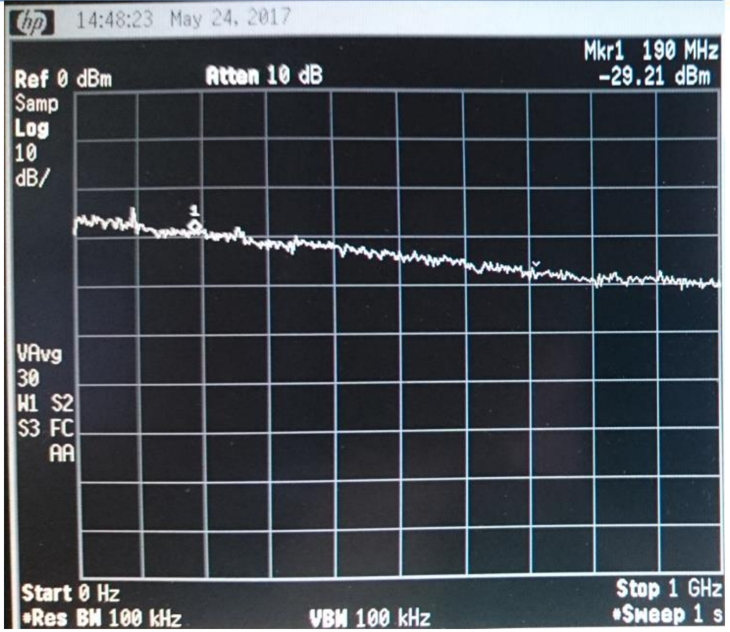


RGB-ITR channel simulator ::NOISE SOURCE

NOISE SOURCE



It is a very low cost noise source (15E on Ebay) with internal amplifiers working in non-linear regime: *pf* values are tool low (nominal *pf=6.0*, measured ***pf=3.2***)



HP E441B, BW=100kHz

$P_N = -30\ dBm = 1.0\ \mu W;$

$p_n = \mathbf{10.0\ pW/Hz}$

TEK 11401, BW = 600MHz

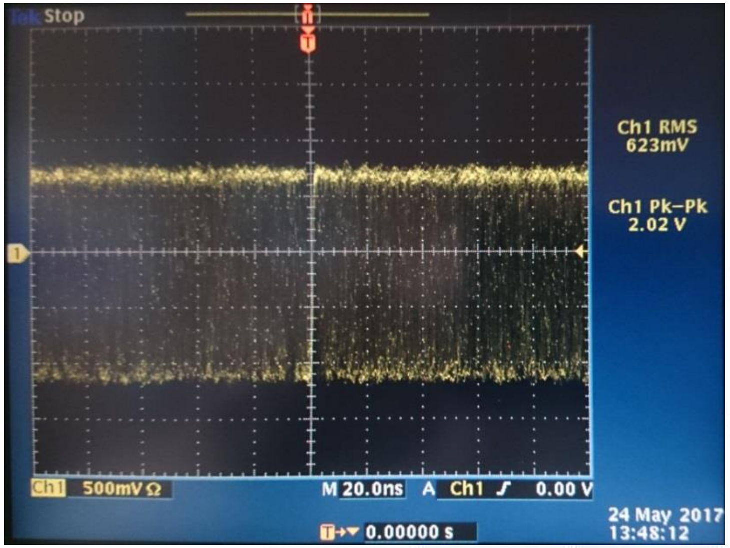
$V_{RMS} = 609.9\ mV;$

$V_{pp} = 2.00\ V;$

$pf = 3.27;$

$P_N = 7.44\ mW = 8.7dBm;$

$p_n = \mathbf{7.89\ pW/Hz}$



TEK 3034, BW = 300MHz

$V_{RMS} = 623mV;\ V_{pp} = 2.02V; pf = 3.24;$

$P_N = 7.76\ mW = 8.9\ dBm; p_n = \mathbf{16.5\ pW/Hz}$

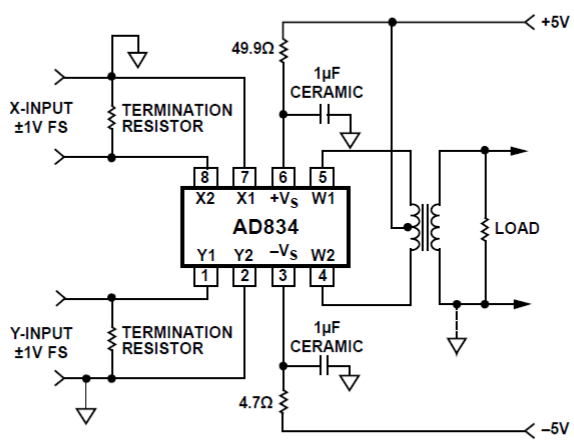
RGB-ITR channel simulator :: AMPLITUDE MODULATOR

An amplitude modulator is used to simulate the variable reflectivity of the target.

The AM mod. has two inputs (the carrier/reference signal REF and the analog modulation signal V_{AM}) and one output $V_{TAM} = REF * V_{AM}$. The AM mod. is designed around the AD 834, DC to 500 MHz, four quadrant multiplier ($W = X * Y$) with:

- Differential +/-1V full scale inputs (X, Y); $P_{IN}(max) = 10\ mW = 10\ dBm$
- Differential +/- 4mA output current (W); $P_{OUT}(max) = 0.4\ mW = -4\ dBm$
- NOTA BENE: the full scale gain of this multiplier is **-14 dB**!
- Low distortion (0.05% for 0 dBm input)

The differential, push-pull current outputs (W1, W2) drive a center-tap (2:1 ratio), wideband (0.4/450 MHz), 50 Ω matched transformer ADT21T (Mini Circuits). The carrier frequency is Fixed at 190 MHz, therefore DC coupling is not mandatory. The symmetric transformer attenuates the odd harmonics of the output signal.



Transformer Electrical Specifications

Ω RATIO (Secondary/Primary)	FREQUENCY (MHz)	INSERTION LOSS*			PHASE UNBALANCE (Deg.) Typ.		AMPLITUDE UNBALANCE (dB) Typ.	
		3 dB MHz	2 dB MHz	1 dB MHz	1 dB bandwidth	2 dB bandwidth	1 dB bandwidth	2 dB bandwidth
2	0.4-450	0.4-450	0.6-400	1-200	1	1	0.2	0.3

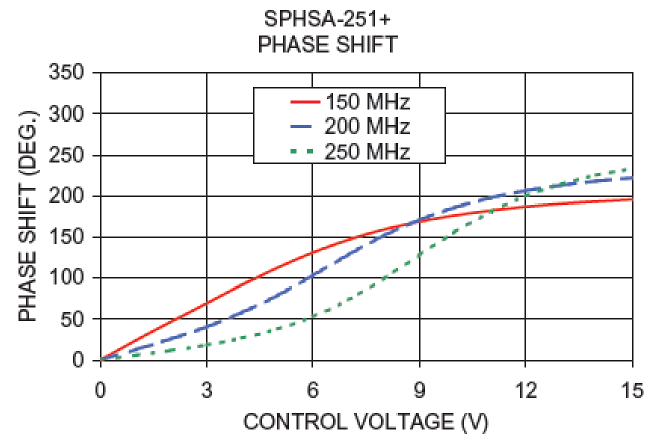
* Insertion Loss is referenced to mid-band loss, 0.4 dB typ.

RGB-ITR channel simulator :: PHASE MODULATOR

A phase modulator is used to simulate the variable shift caused by the distance of the target. The Mini Circuits SPHSA-251+ is a voltage variable phase shifter providing 180° phase control from 150 to 250 MHz; the control bandwidth is DC to 30 kHz and the control voltage is 0 to 15 V. The typical insertion loss is 1.5 dB; the maximum input power is 20 dBm.

Unfortunately the insertion loss and VSWR are not independent from the control voltage. This dependence cause some unwanted residual AM modulation. For the moment this passive phase shifter is the best “ready-made” component available.

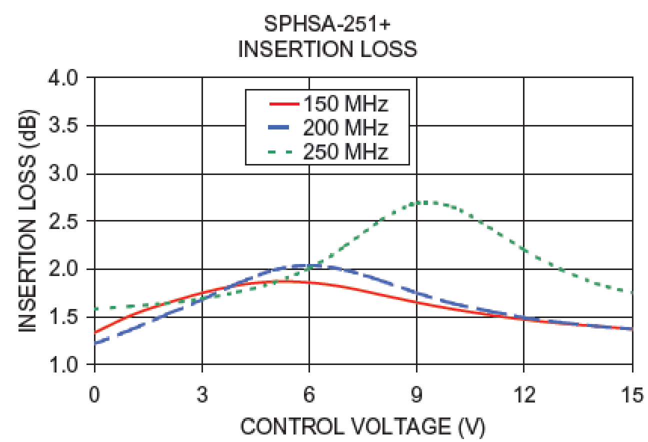
In order to obtain full 0-360° variable phase shift, two devices are used in cascade.



Typical Performance Data

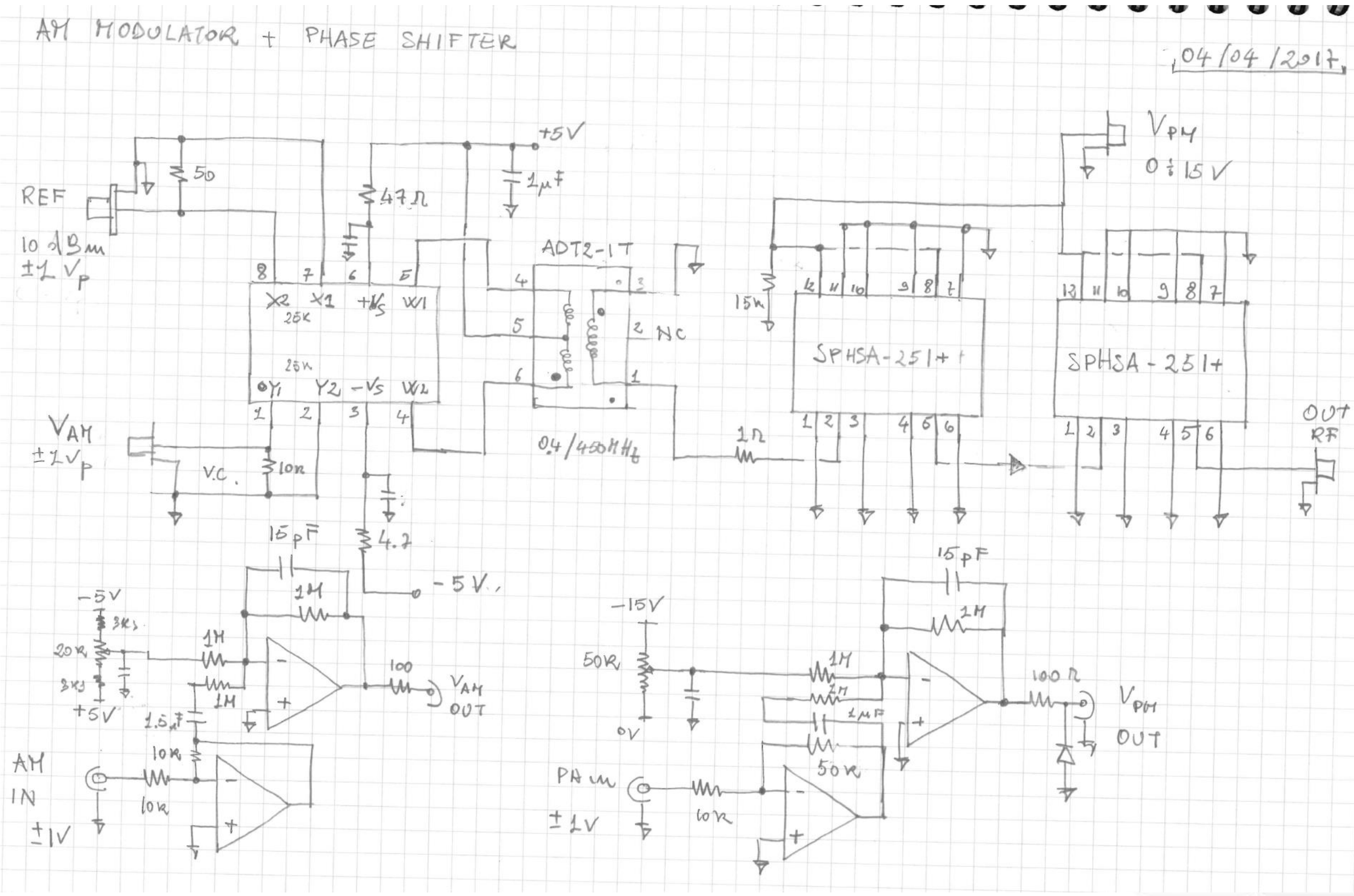
Control Voltage (V)	Phase Shift* (Degrees)			VSWR (:1)			Insertion Loss (dB)		
	150 MHz	200 MHz	250 MHz	150 MHz	200 MHz	250 MHz	150 MHz	200 MHz	250 MHz
0.0	0.01	0.01	0.01	1.10	1.14	1.73	1.33	1.22	1.58
1.0	24.23	12.97	5.88	1.07	1.19	1.70	1.51	1.36	1.61
2.0	46.91	25.98	11.76	1.07	1.25	1.66	1.64	1.52	1.64
3.0	69.53	40.70	18.54	1.17	1.31	1.61	1.75	1.68	1.69
4.0	91.78	58.22	27.01	1.30	1.36	1.53	1.83	1.85	1.76
5.0	112.67	79.19	38.16	1.45	1.38	1.43	1.87	1.99	1.86
6.0	131.26	103.36	53.33	1.58	1.36	1.29	1.86	2.04	2.01
7.0	146.85	128.64	73.76	1.67	1.30	1.13	1.81	1.99	2.24
8.0	159.33	151.94	99.51	1.72	1.25	1.13	1.73	1.88	2.51
9.0	169.05	171.21	128.38	1.74	1.25	1.30	1.65	1.75	2.69
10.0	176.58	186.24	156.49	1.75	1.27	1.41	1.58	1.64	2.65
11.0	182.48	197.74	180.79	1.75	1.31	1.42	1.52	1.56	2.44
12.0	187.09	206.48	200.05	1.75	1.34	1.37	1.47	1.49	2.20
13.0	190.71	213.11	214.63	1.74	1.36	1.31	1.43	1.44	2.00
14.0	193.55	218.17	225.44	1.73	1.37	1.27	1.40	1.40	1.85
15.0	195.79	222.07	233.49	1.73	1.39	1.24	1.37	1.37	1.75

* Normalized at control voltage = 0V



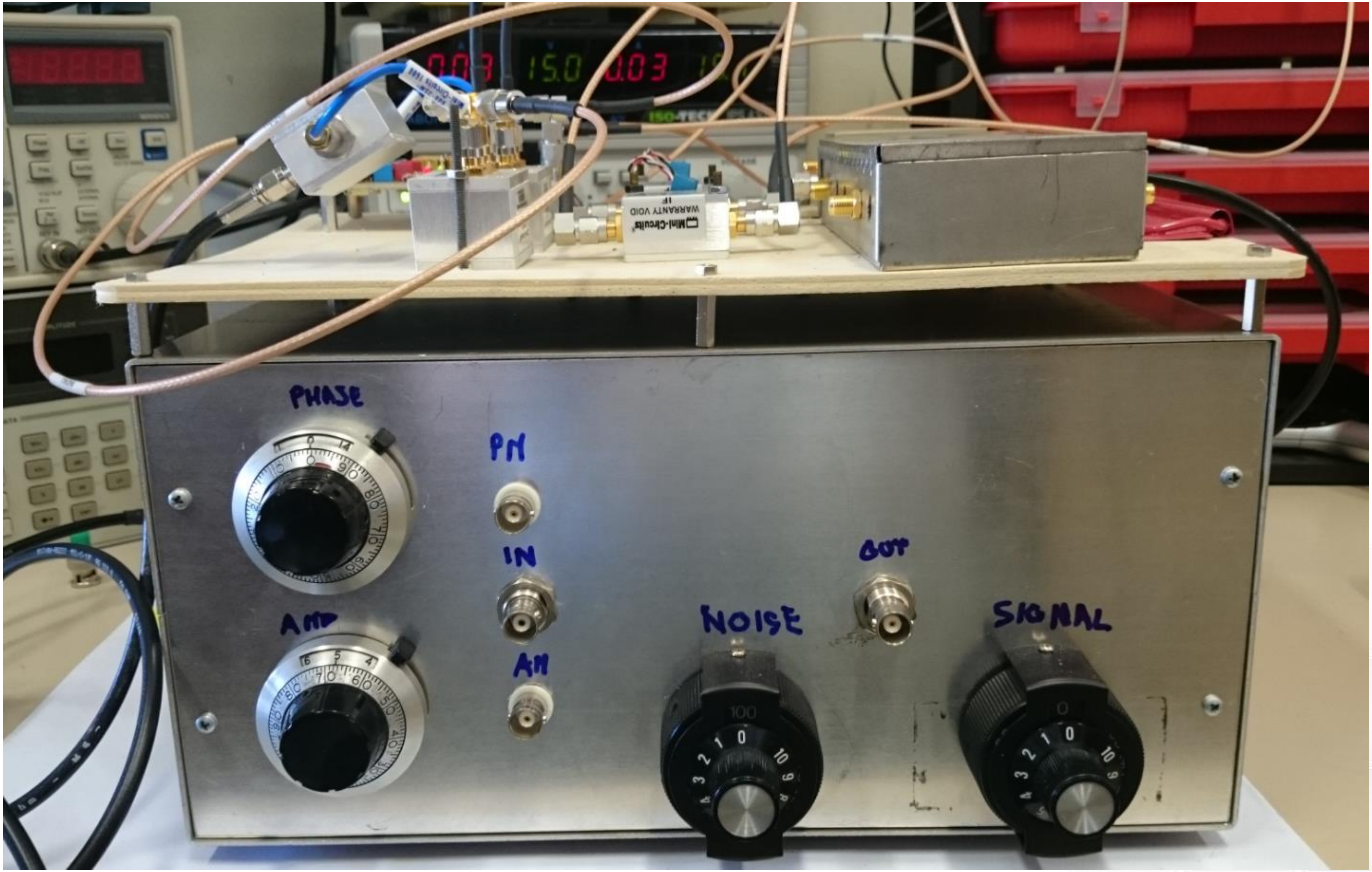
RGB-ITR channel simulator realization

Amplitude and phase modulators.



RGB-ITR channel simulator realization

RGB-ITR simulator control panel.



LOCK-IN AMPLIFIER (LIA):: DEFINITION

An amplifier that can extract a signal with a known carrier wave (F_M) from an *extremely noisy environment*.

Depending on the dynamic reserve ("headroom") of the instrument, signals up to 1 million times smaller *(-120 dB)* than noise components, potentially fairly close by in frequency, can still be reliably detected and measured (absolute amplitude and phase respect a reference).

The LIA is a homodyne detector (*synchronous demodulator*) followed by *low pass filter* adjustable in cut off frequency (time constant) and filter order (frequency slope).

Source: *Wikipedia*.

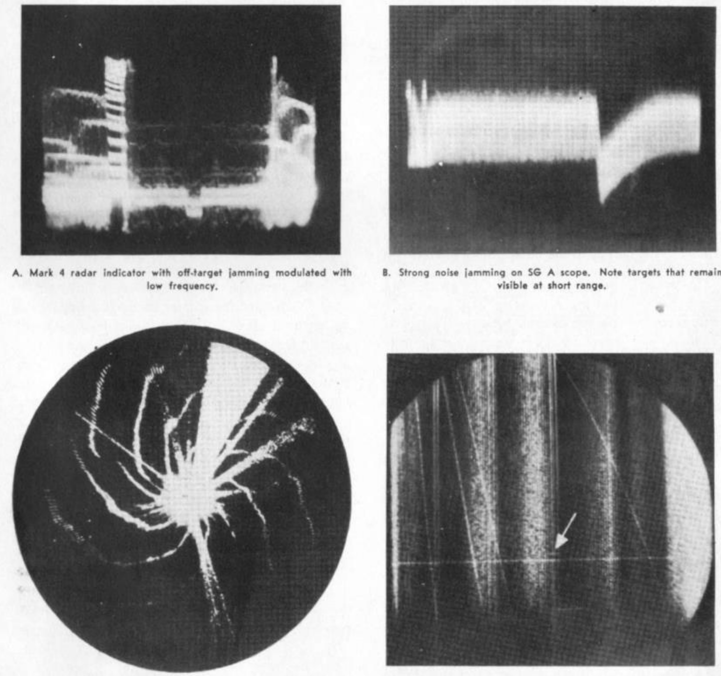
LIA :: (almost) nothing new under the sun

Coherent demodulation theory is based on the mathematical theory developed by Jean Baptiste Joseph Fourier (Fourier series, 1830).

First practical implementations begin in the 19th century (electro-mechanical and vacuum tube circuits). With solid state electronics, the LIA becomes a standard and commercially available instruments.

RADAR use electronic circuits to process naturally or artificially (electronic counter measures, ECM) noisy signals. RADAR ECM was invented in 1940 by German; ECM is evolved during the Cold War with a peak during the Vietnam conflict, when HP introduces the panoramic, (calibrated in dBm) spectrum analyzer in 1969.

Coherent demodulation is the typical architecture of modern RADAR receivers.



A. Mark 4 radar indicator with off-target jamming modulated with low frequency.

B. Strong noise jamming on SG A scope. Note targets that remain visible at short range.

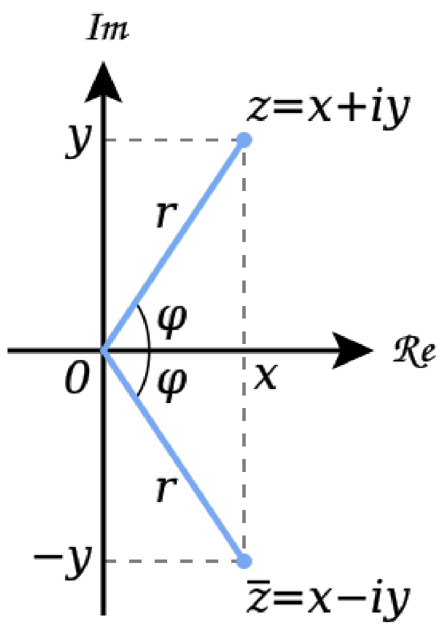
C. Strong noise jamming on PPI. Note pick-up in side and back lobes. Spiral lines are due to interference from other radar.

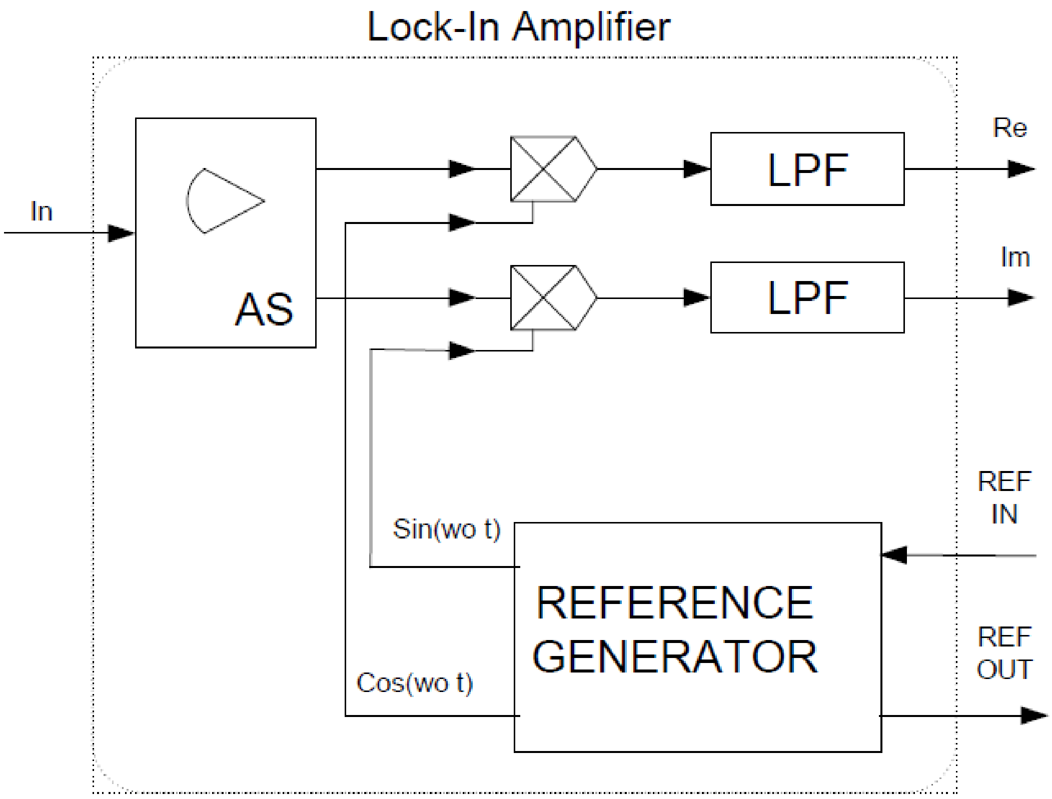
D. Mark 8 B scope partially jammed with noise modulated jamming. Note echo still visible at arrow.

Theory and practice of LIA

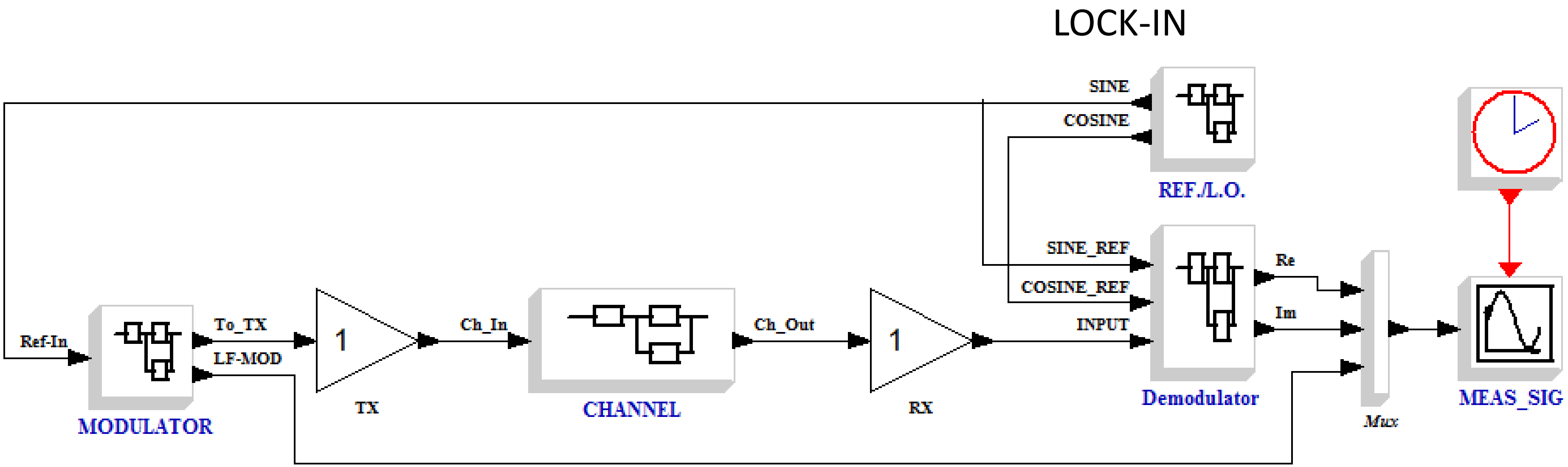
$$Re = \frac{1}{T}\int_{-T/2}^{T/2} In(t) \; * Cos(\omega_M t)\, dt$$

$$Im = \frac{1}{T}\int_{-T/2}^{T/2} In(t) \; * Sin(\omega_M t) dt$$





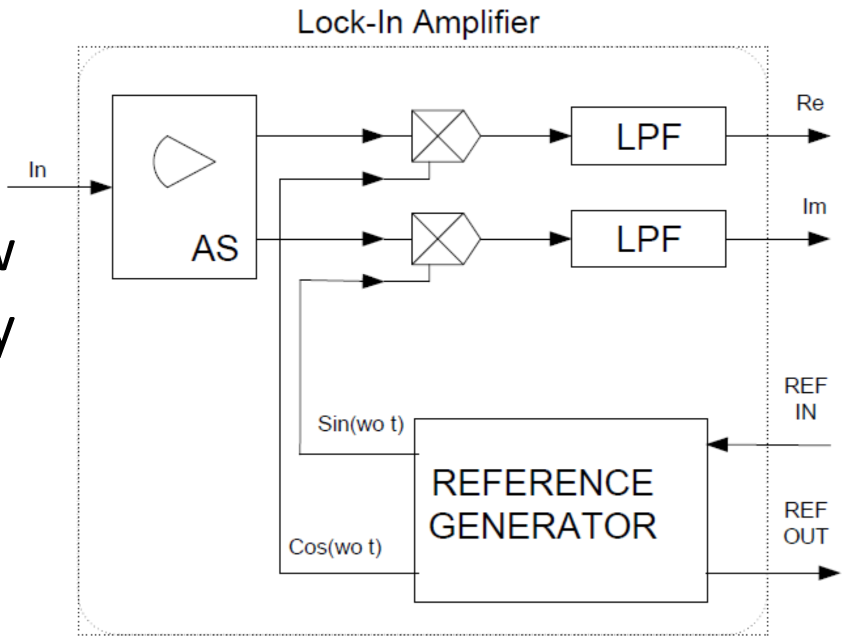
Typical application diagram



In this configuration the PLL section is not required because the carrier signal is directly modulated by the *internal* SINE reference (**Ref-In**).

Step-by-step design procedure

- **AS**: Amplifier/Splitter. A device with an input and two outputs. The two output bust be an exact copy of the input signal (except amplitude change and fixed [transit] time delay.
- **REF.GEN.** A sinusoidal generator, fixed frequency, fixed amplitude; the two outputs have same amplitude but 90° relative phase
- **Multiplier/Mixer**: active or passive component capable to implement (in analog current/voltage) the multiplication of two signals.
- **Low Pass Filter.** Programmable low pass filter (order and frequency corner).



How to select the physical components?

Two key parameters: **bandwidth** and **dynamic range**

Bandwidth: F_{min} and F_{max} of the device, usually expressed in term of -3 dB frequency. In our application the real signal (information) bandwidth is limited at +/- 10 kHz around a 190 MHz carrier, therefore

$$F_{min} < F_M/2 \; ; F_{max} > 2 * F_M$$

Dynamic range: *maximum signal / minimum signal*.
Maximum signal: the biggest signal that I can put at the input and the device is still operate in the “linear region” (no saturation).
Minimum signal (noise floor): the smallest signal at the input still usable (measurable) despite the noise.

How to select the physical components?

Maximum signal: the biggest signal that I can put at the input and the device still operate in the "linear region".

Sinusoidal Signal
V(RMS) = 1
V(peak) = 1.41
V(p-p) = 2.82

Noise Signal (wideband, Gaussian)
V(RMS) = 1
V(peak) = 3 (99.7%)
V(p-p) = 6 (99.7%)

In term of maximum allowable signal power, the noise (and we have a very noisy input signals) is A LOT more difficult to handle than sinusoidal signals.

Despite the very conservative "headroom" (+/- 3 sigma cover of 99.7% of possible instantaneous amplitudes), a systematic measurement error is still present.

What happens if I overload the input?

The final behavior of the instrument depend by many factors, therefore useful model of performance degradation can be developed only by experimental procedures and detailed numerical simulations.

The performance degradation (systematic measurement errors) is progressive. Usually the LIA includes some overload detection circuits along the signal path.

What happens if the signal is too low?

Minimum signal (noise floor): the smallest signal at the input still usable (measurable) despite the noise.

If the input signal (that carry information) have an amplitude too low respect the “noise floor”, the final measure will be afflicted by noise at the point to be useless.

F.U.B.A.R.

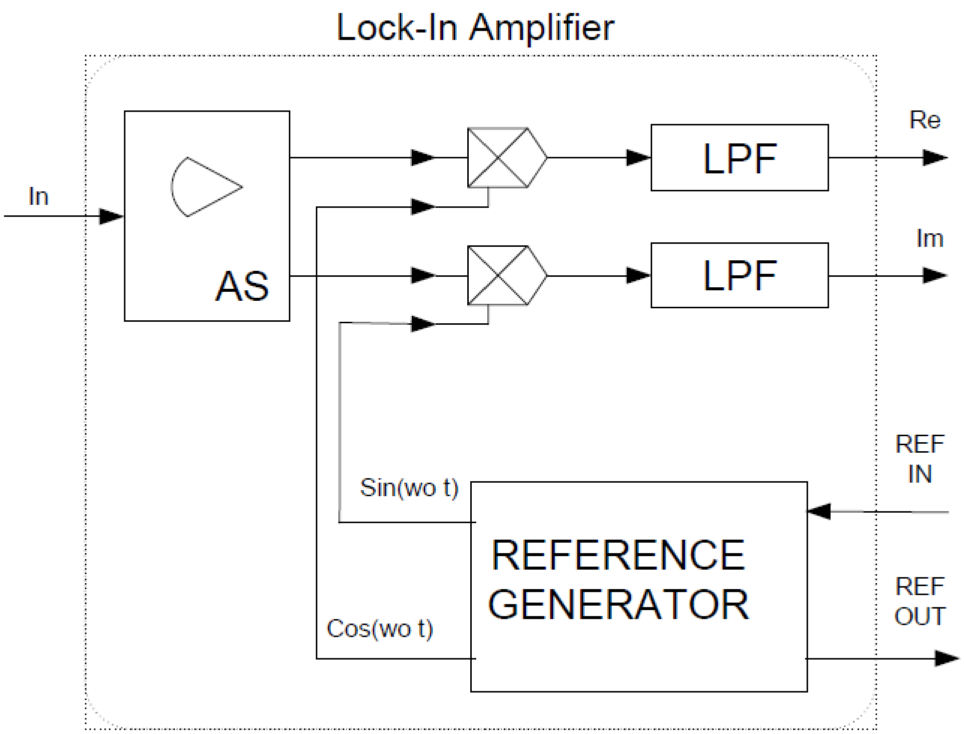
Fucked Up Beyond Any Repair

If the signal (plus noise) is too high, the components exit from the “linear region” (they saturate) and the measure is **FUBAR**.

If the signal (respect the noise) is too low, the recovered signal is still affects by noise and the measurement is **FUBAR**.

Design a LIA (physical component selection) is a constant fight between bandwidth and dynamic range matching between interconnected components.

What is the most critical component of an analog LIA?

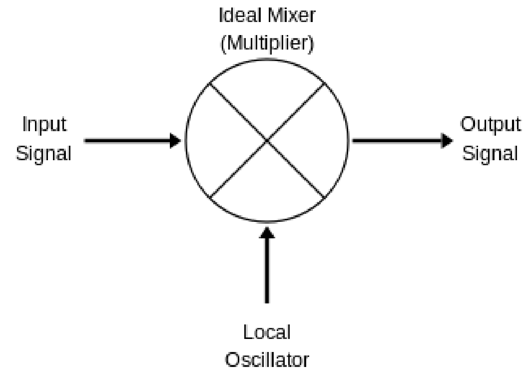


1. **AS**: Amplifier/Splitter. A device with an input and two outputs. The two output bust be an exact copy of the input signal (except amplitude change and fixed [transit] time delay.
2. **REF.GEN**. A sinusoidal generator, fixed frequency, fixed amplitude; the two outputs have same amplitude but 90° relative phase
3. **Multiplier/Mixer**: active or passive component capable to implement (in analog current/voltage) the multiplication of two signals
4. **LPF**: (low pass filter) an active filter that sets the effective bandwidth of the recovered signal

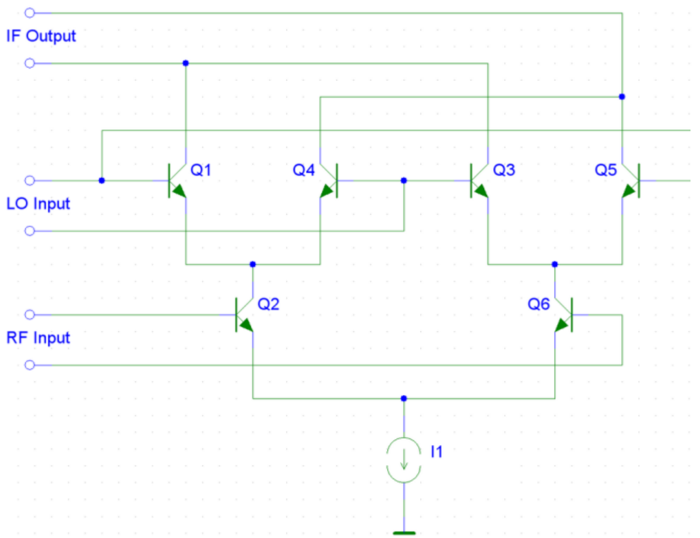
The most critical component of an analog LIA is the analog multiplier

The design of an analog LIA must start from the analog multiplier.
Basically, there are two technologies:

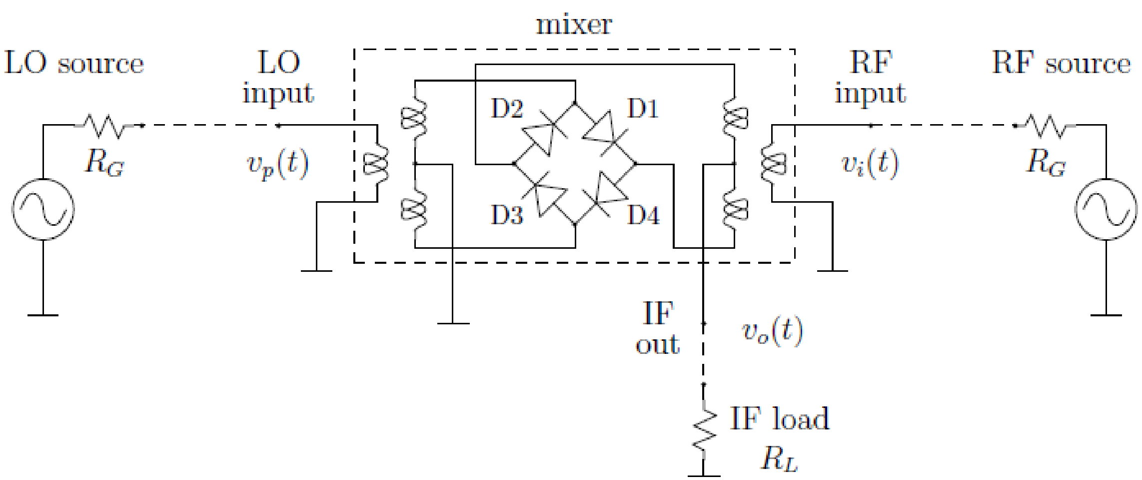
- **Active analog multiplier**, using Gilbert cell integrated circuits
- **Passive analog mixers**, using transformers and diodes



Gilbert cell multiplier



Double balanced diode mixer

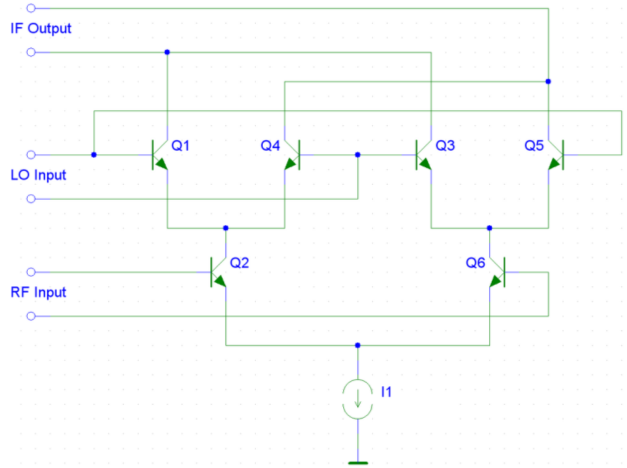


Gilbert cell analog multiplier

The trans conductance gain ($g_m = \frac{i_c}{v_b} = I_C/V_T$) of an (ideal) bipolar junction transistor depends by the polarization current I_C. The polarization current can be used as modulating input (multiplier). Invented by Barry Gilbert working in Tektronix (1968) in order to realize continuous variable gain with 1-10 range with constant bandwidth.

Advantages:
- Works from DC ($F_{min} = 0$)
- Very high precision at low freq.
- Fully solid state
- Normalized in/out voltage/current

Disadvantages:
- Limited bandwidth (500 MHz)
- Limited precision (phase distortion) at high frequency
- Temperature dependent DC offset



Double balanced diode mixer as analog multiplier

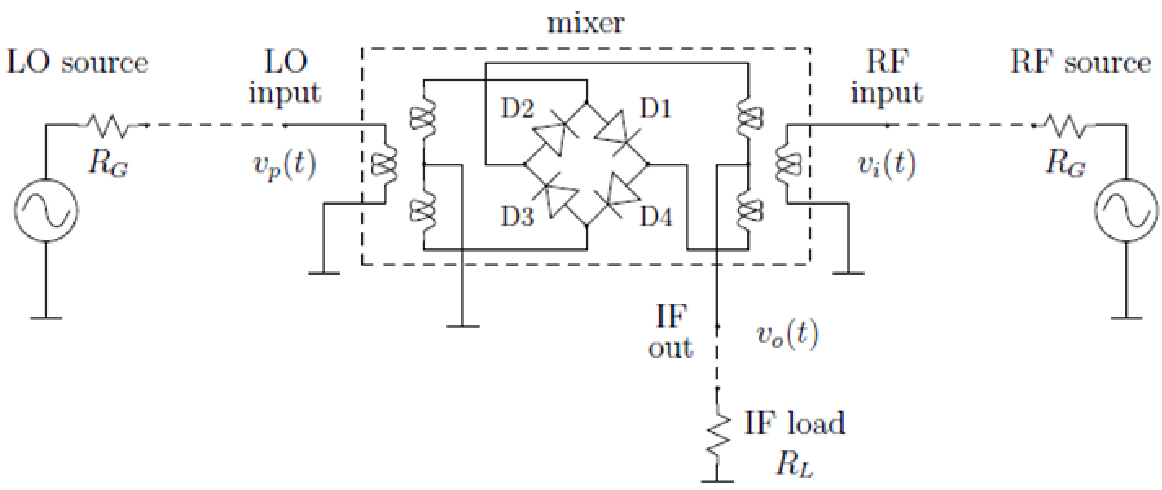
The double balanced diode mixer is a device invented for the first super-heterodyne (frequency conversion) radios (1910, still vacuum tube technology) later optimized by HP for high accuracy radio frequency measurement systems.

Advantages:

- Very high precision
- Very high dynamic range (>60 dB)
- Very high F_{max} (GHz)

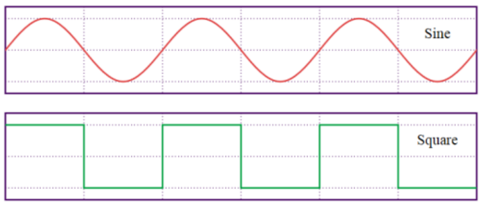
Disadvantages:

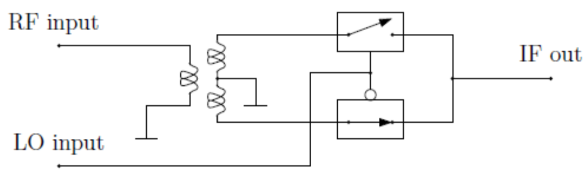
- F_{min} is not zero (transformer coupling)
- Harmonic mixing
- Limited port isolation
- Temperature dependent DC offset
- Easy to damage permanently using excessive power at the input ports (LO and RF)

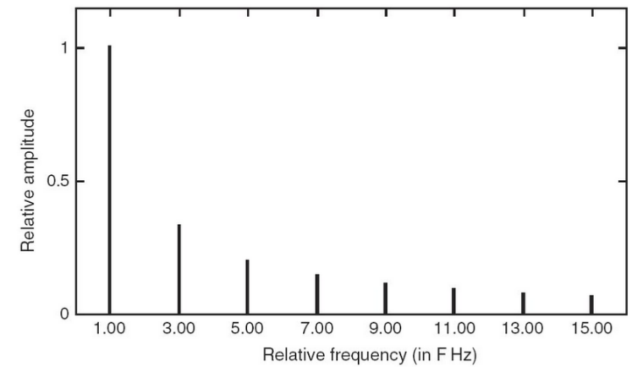


Double balanced diode mixer as analog multiplier

A double balanced mixer does not implement a REAL analog multiplication. The LO input signal is used to open and close (alternatively) two diodes of the bridge implementing a +1/-1 multiplication of the RF input signal with the equivalent SQUARE WAVE signal on the LO (REF) input port.







Mathematically speaking, it is how to use an ***ideal*** analog multiplier with the (LO) input connected with a SQUARE WAVE reference generator of frequency ω_R.
The square wave can be expanded using Fourier series as

$$sqwf(t) = \sin(\omega_R t) + \left[\frac{1}{3}\right]\sin(3\omega_R t) + \left[\frac{1}{5}\right]\sin(5\omega_R t) + \left[\frac{1}{7}\right]\sin(7\omega_R t) + \cdots$$

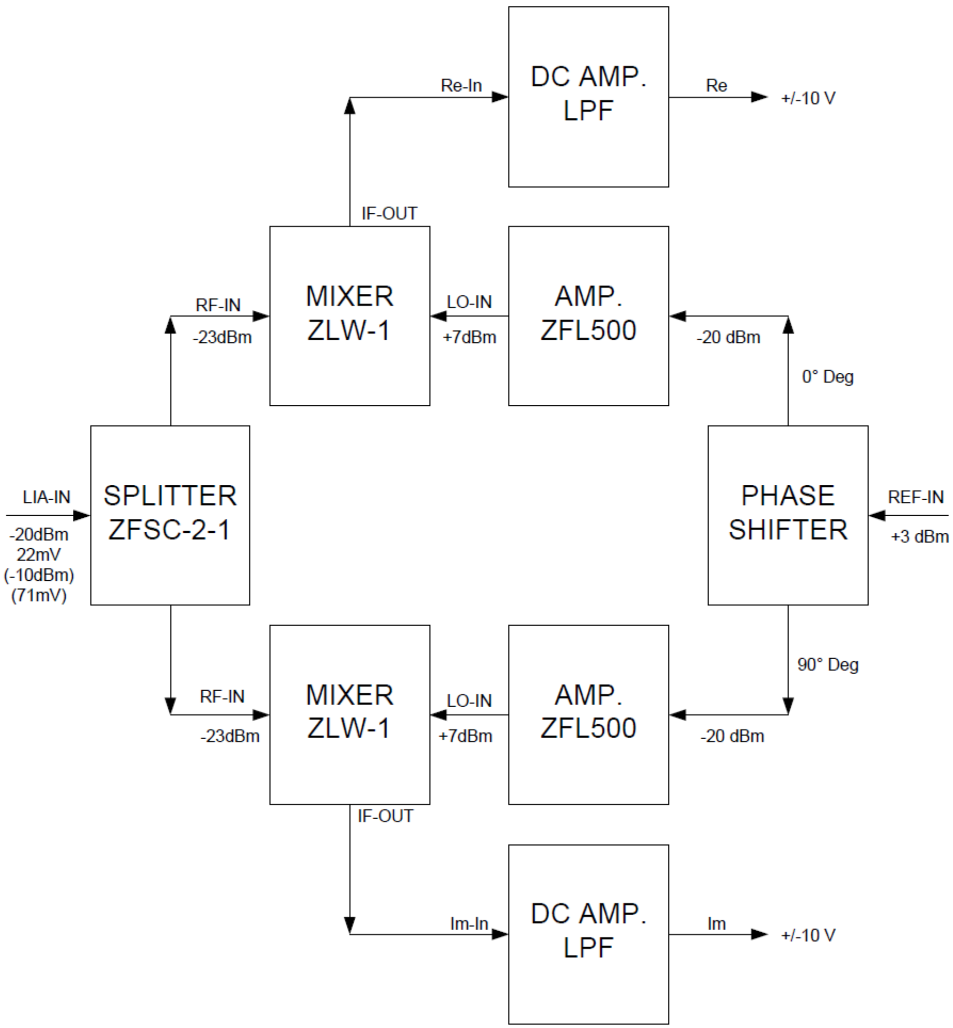
This "collateral damage" is called "harmonic demodulation": the IF output signal depends not only by $F_R(\omega_R)$but also by its harmonics. The harmonic demodulation may introduce a systematic error in the measurement (for more details: SR844 Stanford Research manual).

God Bless Mini Circuits

Design a build from scratch a double balanced diode mixer (and other components) with GHz bandwidth is a full time job that demand very specific expertize, test instruments, access to very special electronic components and a lot of time. A better solution is to chose from a catalog of ready and tested component.

For radio frequency circuits, Mini Circuits is the best choice.

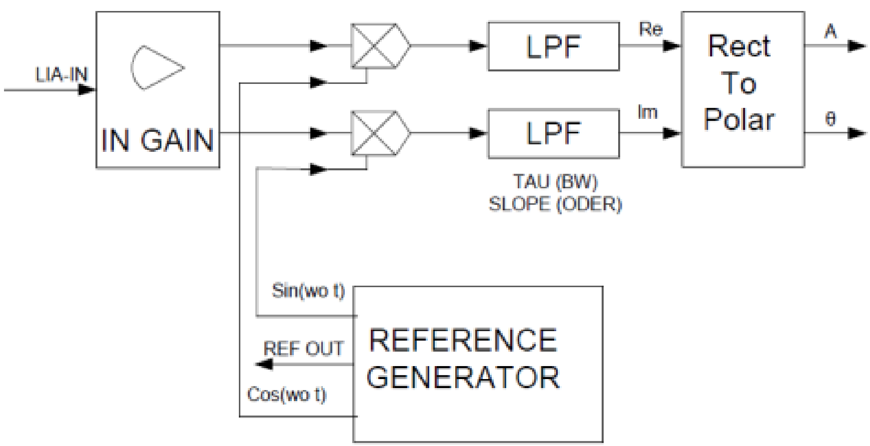
Lock-In Amplifier Schematic Diagram



$$Re = \frac{1}{T}\int_{-T/2}^{T/2} In(t) \ * Cos(\omega_R t) \ dt$$

$$Im = \frac{1}{T}\int_{-T/2}^{T/2} In(t) \ * Sin(\omega_R t) dt$$

The reference signal comes from the HP 5687A signal generator.



Signal Splitter

Each mixer need a copy of the input signal. With RF signals, put two RF-IN in parallel is a really bad idea, especially inside a measurement system.

ZFSC-2-1+

BNC version shown

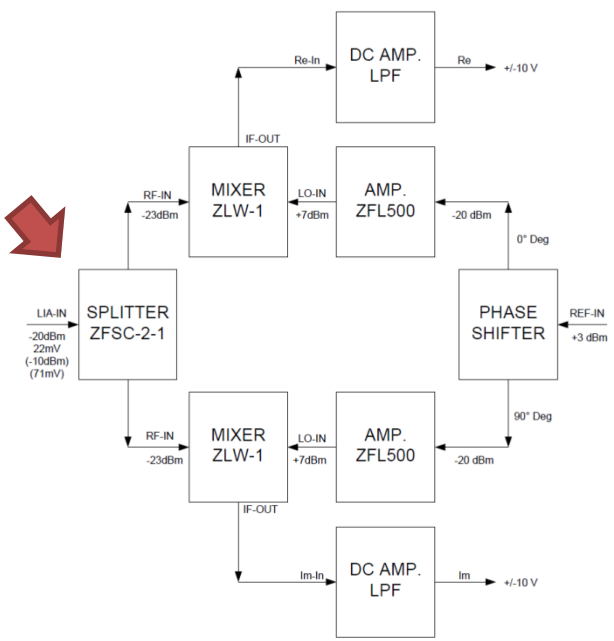
Features
- wideband, 5 to 500 MHz
- low insertion loss, 0.3 dB typ.
- excellent isolation, 28 dB typ.
- excellent amplitude unbalance, 0.1 dB typ.
- good VSWR, 1.2:1 typ.
- rugged shielded case

Applications
- VHF/UHF
- instrumentation
- communication systems

Electrical Specifications

FREQ. RANGE (MHz)	ISOLATION (dB)						INSERTION LOSS (dB) ABOVE 3.0 dB						PHASE UNBALANCE (Degrees)			AMPLITUDE UNBALANCE (dB)		
	L		M		U		L		M		U		L	M	U	L	M	U
f_L-f_U	Typ.	Min	Typ.	Min	Typ.	Min	Typ.	Max.	Typ.	Max.	Typ.	Max.	Max.	Max.	Max.	Max.	Max.	Max.
5-500	30	25	28	20	25	20	0.2	0.5	0.3	0.6	0.6	0.8	2	4	4	0.15	0.15	0.30

L = low range [f_L to 10 f_L] M = mid range [10 f_L to f_U/2] U= upper range [f_U/2 to f_U]



In a 50-ohm system, each output would be connected to a 50-ohm impedance, thus offering a 25-ohm impedance to the input port. Thus, the impedance looking into the common or input port would present a mismatch in a 50-ohm system. To correct this mismatch, a 25 to 50-ohm matching transformer would be necessary as shown in Fig. 4.

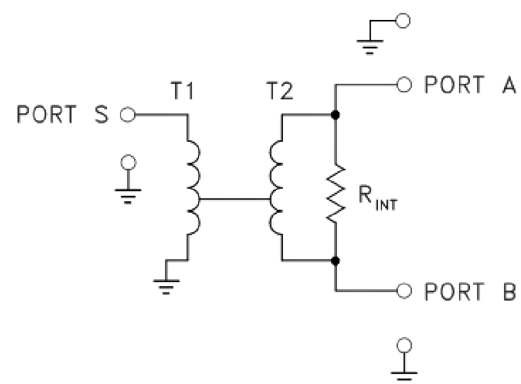


Fig. 4. T1 is a 2:1 impedance matching transformer in the input circuit of the power splitter/combiner.

Mixer

ZLW-1

CASE STYLE: M21

Maximum Ratings

Operating Temperature	-55°C to 100°C
Storage Temperature	-55°C to 100°C
RF Power	50mW
IF Current	40mA

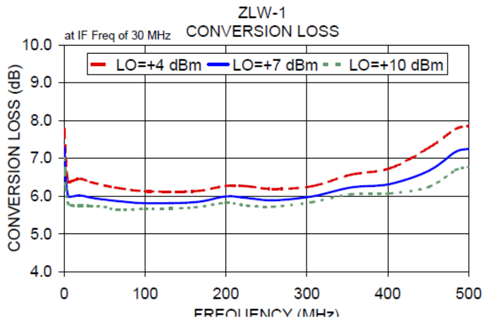
Permanent damage may occur if any of these limits are exceeded.

Features

- low conversion loss, 5.81 dB typ.
- high L-R isolation, 45 dB typ., L-I, 40 dB typ.
- rugged shielded case

Applications

- VHF/UHF
- defense & federal communications
- instrumentations

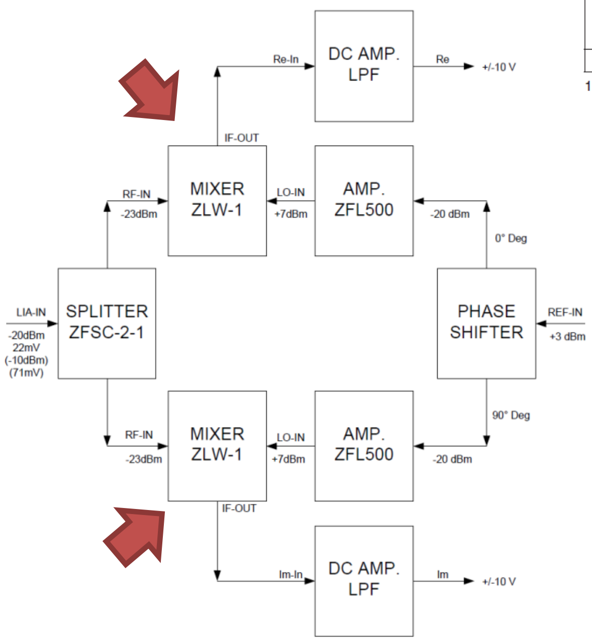


Electrical Specifications

FREQUENCY (MHz)		CONVERSION LOSS (dB)				LO-RF ISOLATION (dB)						LO-IF ISOLATION (dB)					
LO/RF	IF	Mid-Band m			Total Range	L		M		U		L		M		U	
f_L-f_U		$\overline{X}$	σ	Max.	Max.	Typ.	Min.	Typ.	Min.	Typ.	Min.	Typ.	Min.	Typ.	Min.	Typ.	Min.
0.5-500	DC-500	5.81	0.08	7.0	8.5	50	45	45	30	35	25	45	35	40	25	30	20

1 dB COMP.: +1 dBm typ.

L = low range [f_L to 10 f_L] M = mid range [10 f_L to f_U/2] U = upper range [f_U/2 to f_U]
m= mid band [2f_L to f_U/2]



Correct usage of a double balanced RF mixer demands a right combination of art and science:

- RF input power must be limited at 50 mW;
- LO input power must be limited at 10 dBm (10 mW); optimal response is obtained with +7dBm input power from a matched source;
- IF output port must be closed on 50Ω matched load from DC.

The design has been drawn taking in account the previous constraints.

Mixer Driver

The LO input port is driven by a dedicated low noise amplifier. We choose this model because: (1) its low noise and (2) the maximum power output is not capable to damage the mixer.

ZFL-500LN+

SMA version shown

Features

- very low noise, 2.9 dB typ.
- good VSWR, 1.5 :1 typ.
- protected by US Patent, 6,

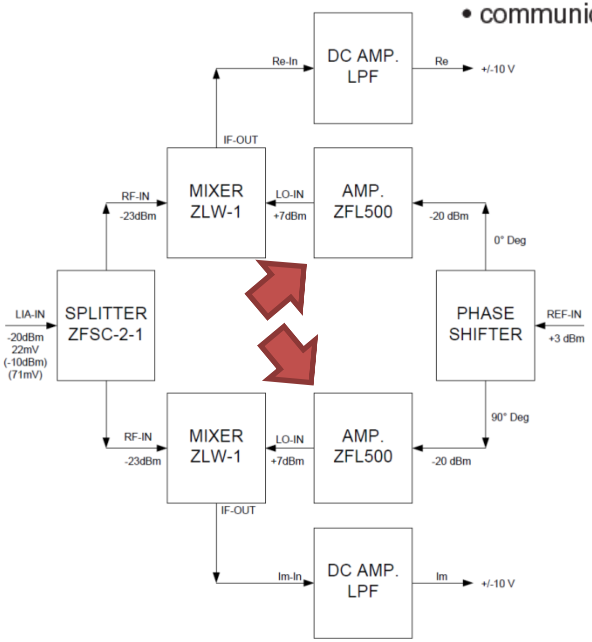
Applications

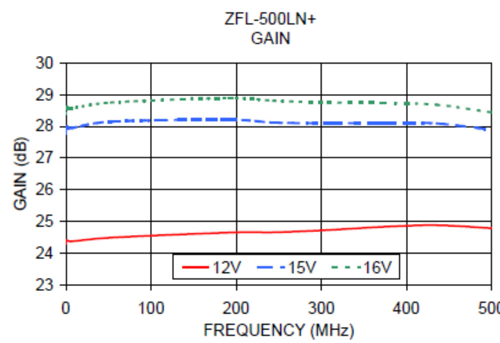
- VHF/UHF
- small signal amplifier
- communications system

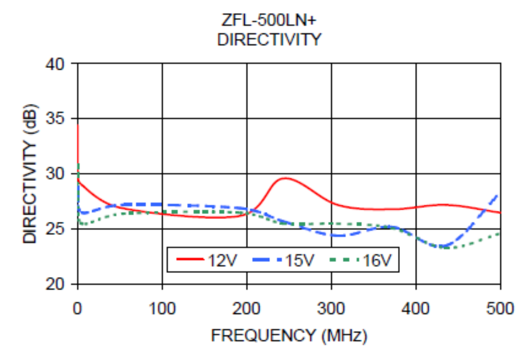
Low Noise Amplifier Electrical Specifications

MODEL NO.	FREQUENCY (MHz) f_L	FREQUENCY (MHz) f_U	NOISE FIGURE (dB) Typ.	GAIN (dB) Min.	GAIN (dB) Flatness Max. Total Range	MAXIMUM POWER (dBm) Output (1 dB Compr.)	MAXIMUM POWER (dBm) Input (no damage)	INTERCEPT POINT (dBm) IP3 Typ.	VSWR (:1) Typ. In	VSWR (:1) Typ. Out	DC POWER Volt (V) Nom.	DC POWER Current (mA) Max.
ZFL-500LN+	0.1	500	2.9	24	±0.5	+5	+5	+14	1.5*	1.6	15	60

m = mid range [2 fL to fU/2]







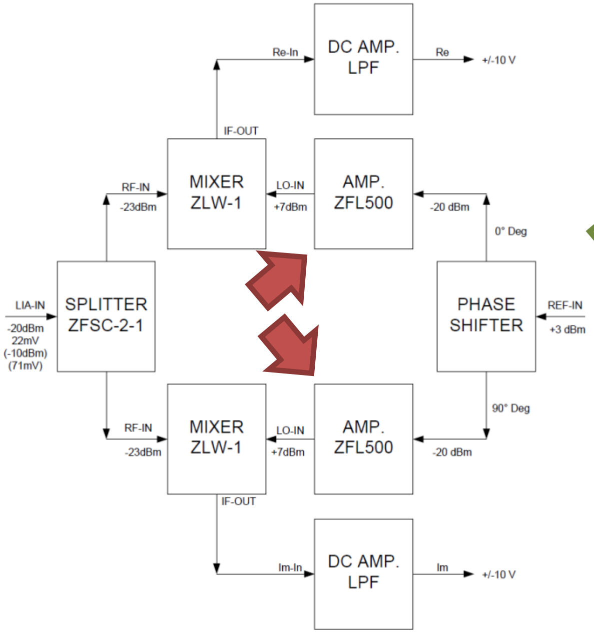
Gain depends by the power supply, therefore each amplifier has a dedicated 12-16 V adjustable power supply.

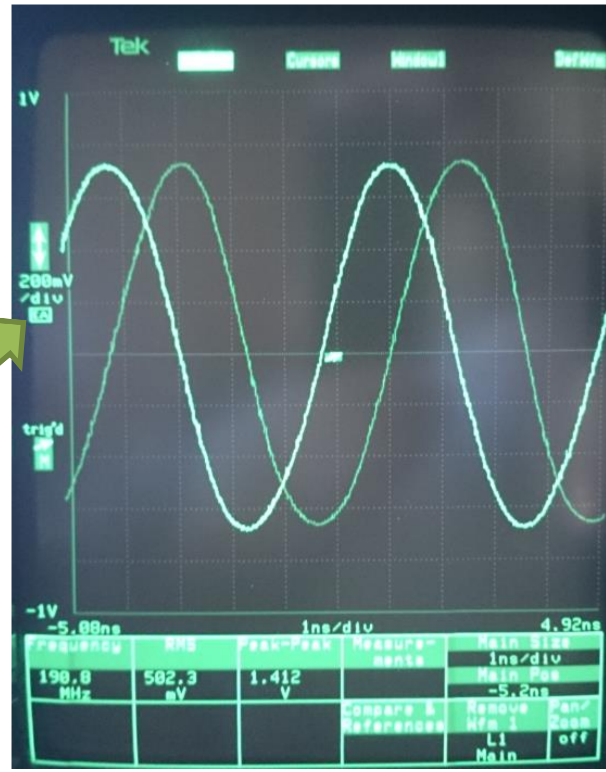
Gain (and phase) has ben adjusted using an high precision, high frequency oscilloscope Tektronix 11401 (600 MHz BW)

Mixer Driver Calibration

ENEA AGENZIA NAZIONALE PER LE NUOVE TECNOLOGIE, L'ENERGIA E LO SVILUPPO ECONOMICO SOSTENIBILE

The gain has been adjusted using individually adjustable power supply.





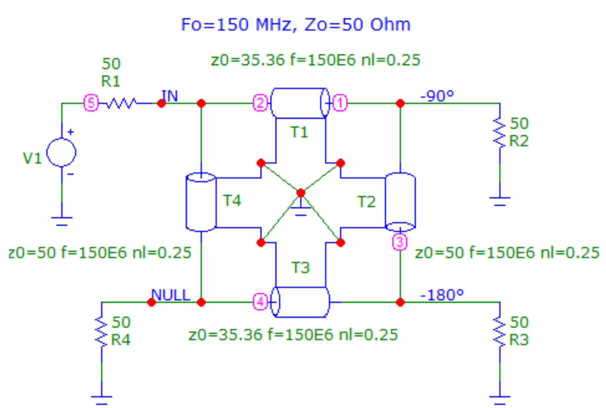
(center) R_E (in phase) channel; (right) I_M (in quadrature) channel. The amplitudes are (in terms of RMS voltage) practically equals (500 mV = +7dBm). The two waveforms are exactly in quadrature. [EA] (Extended Accuracy] means 10 bit accuracy.

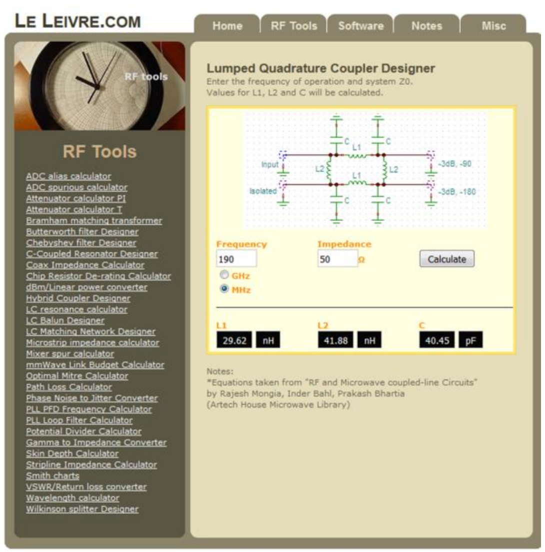
Correct phase has been achieved tuning the phase shifter.

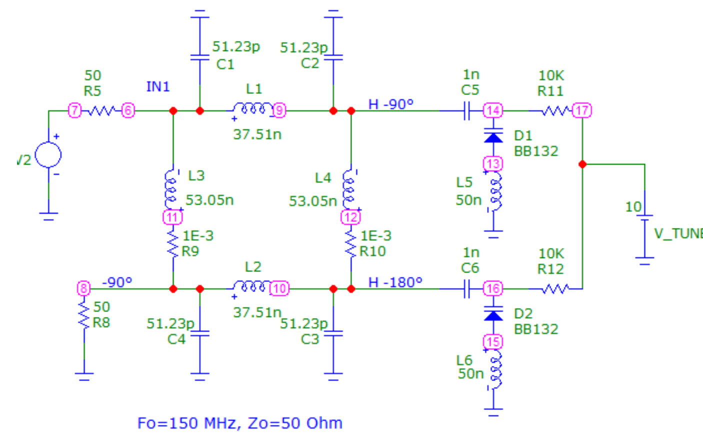
PHASE SHIFTER

The 90° phase shifter has been designed when the variables phase shifter from Mini Circuits was not yet available. A variable phase shifter is like a section of a lossless transmission line where you can tune the length (transit time, phase) using an external parameter (e.g. a voltage).

The best performances are obtained using a 4-ports hybrid coupler with two ports closed on series resonant circuits. Unfortunately, realize custom impedance transmission lines at 190MHz is too cumbersome, therefore, a lumped design (using discrete Ls and Cs) has been chosen.





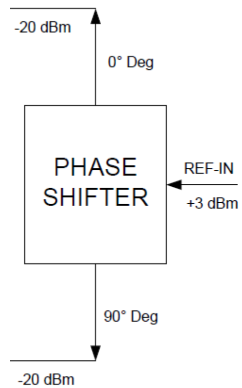


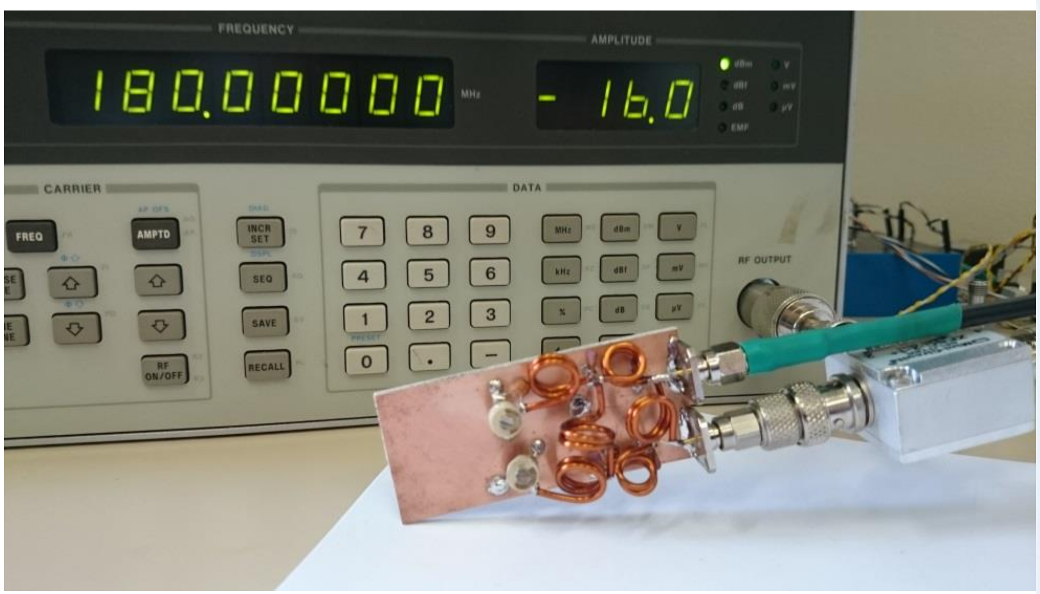
Ref: http://www.qsl.net/yt1vp/50%20MHz%20Noise%20Canceling.htm

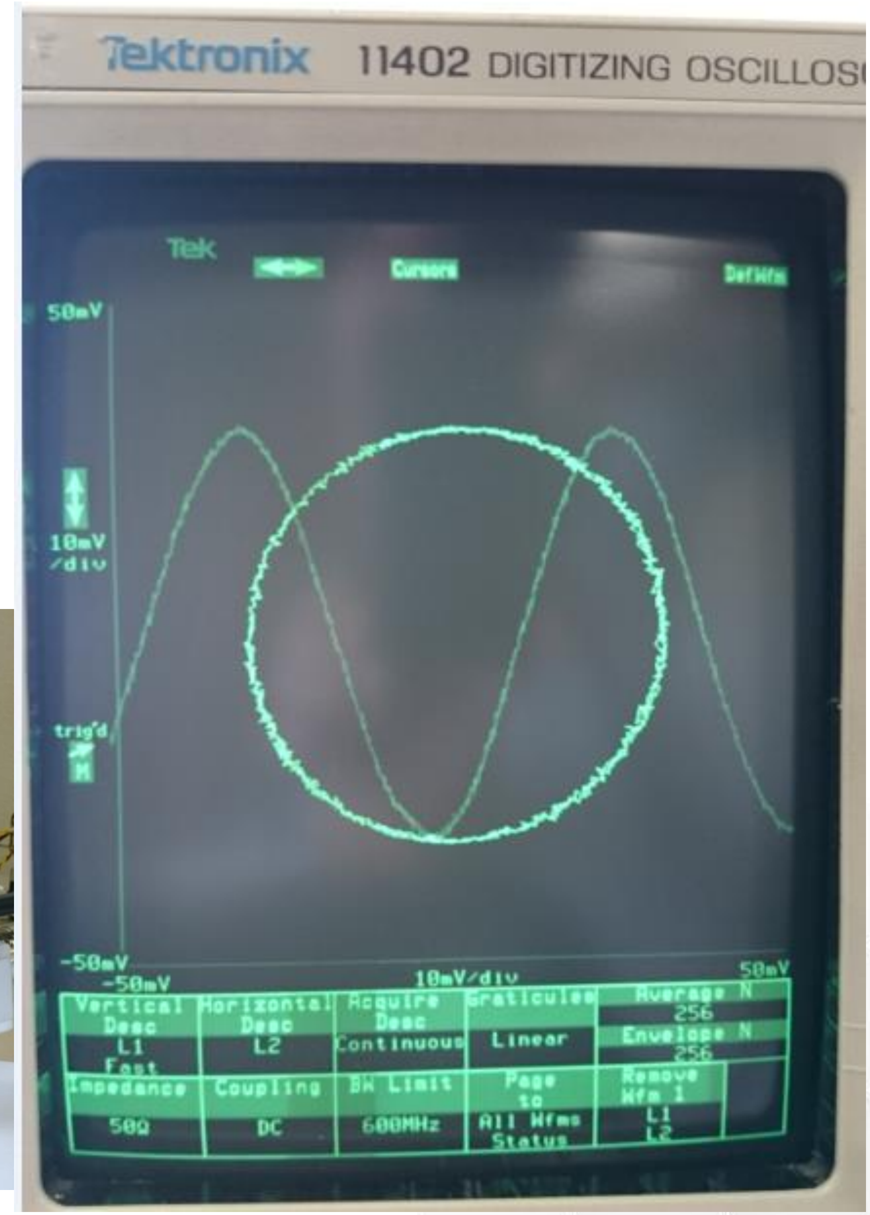
Phase Shifter First Prototype

Reference generator (180 MHz).

The photos shown the first prototype working at 180 MHz. A second circuits has been optimized for 190 MHz.

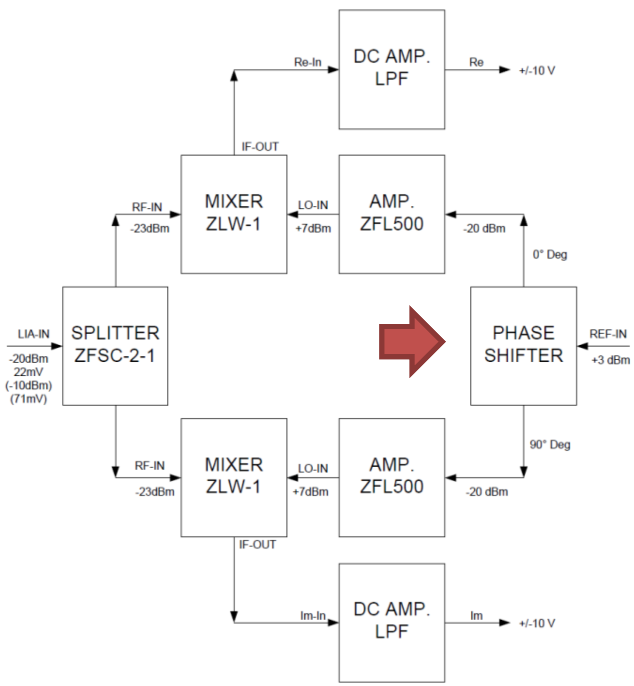


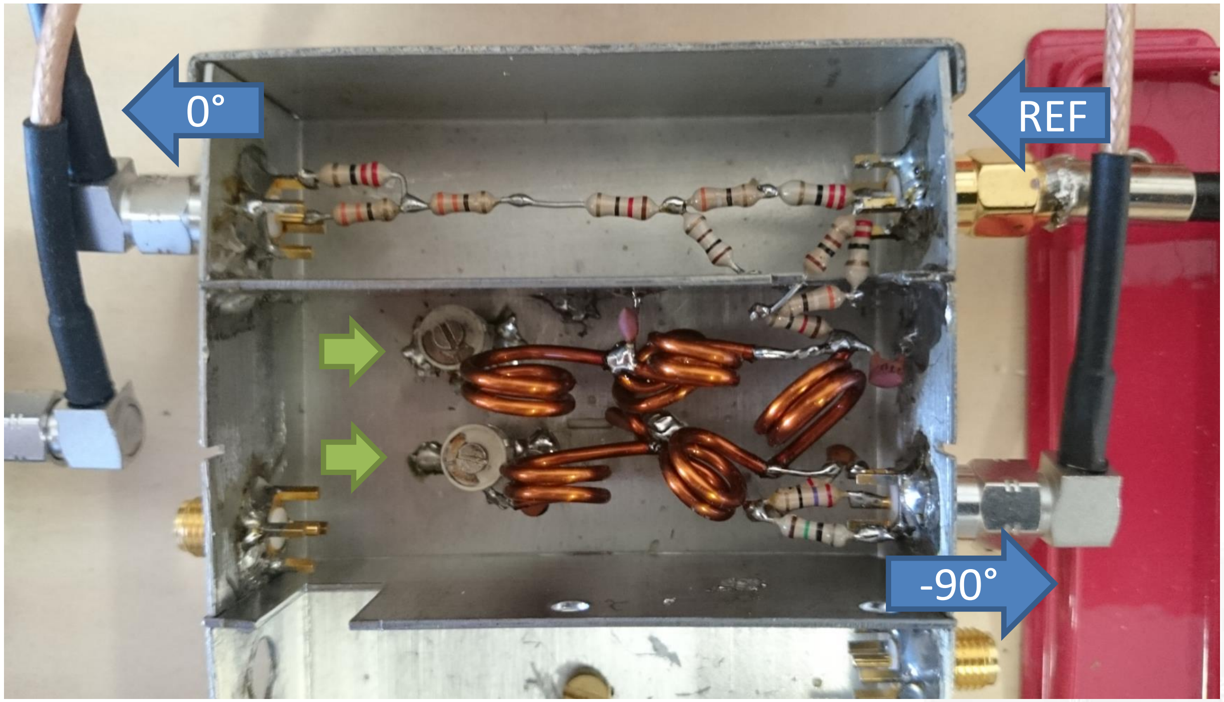




LIA PHASE SHIFTER

The final phase shifter integrate a passive signal splitter (that attenuate the input signal in order to compensate for the fixed gain of the following amplifier) and the tunable quadrature hybrid. Most of the shielded box is still empty: it is ready for the stand-along PLL signal generator that will replace the HP 8657A.



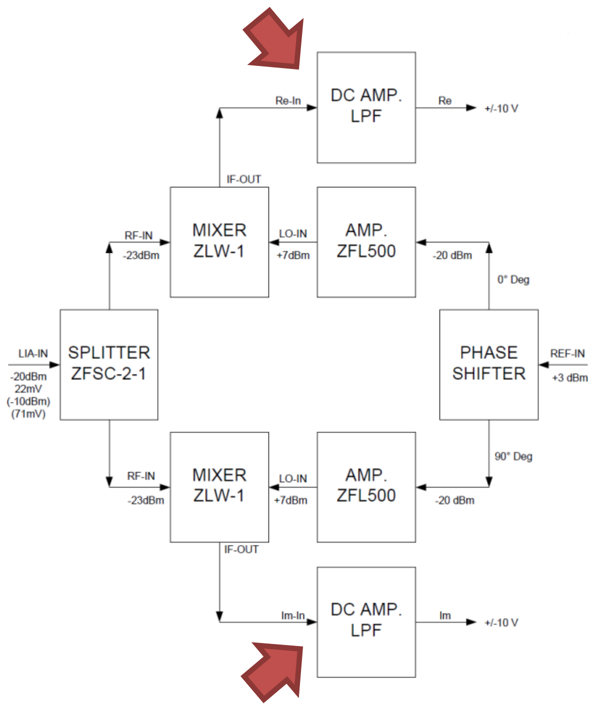


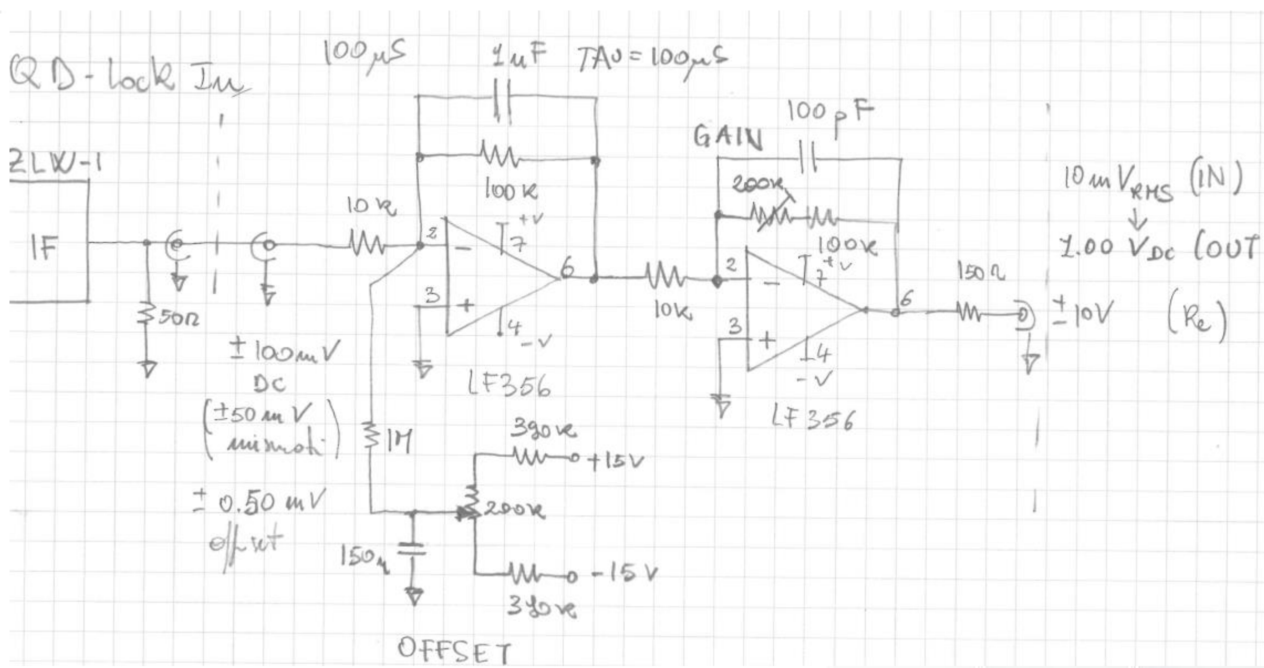
Phase has ben adjusted (green arrows) using an high precision, high frequency oscilloscope Tektronix 11401 (600 MHz BW).

DC AMP + LPF

The mV-level output of the IF mixer ports must be filter with a precise time constants (100μs) and normalized to +/-10 Volts.
The two circuits (Re and Im path) are exactly equals.

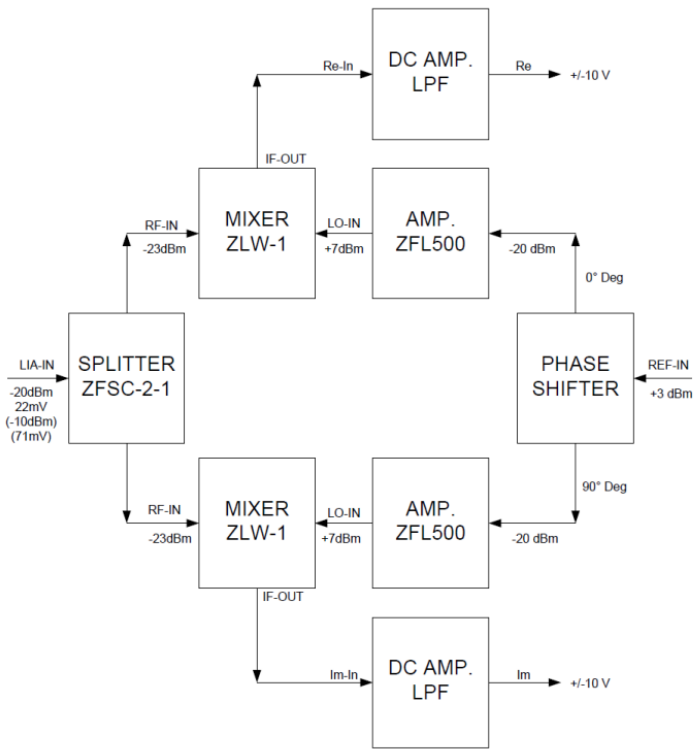
A first inverting stage amplify 10 times and impose a fixed time constant. The first stage includes an external offset compensation. The second stage invert again the signal and finalize the +/- 10 Volt output.

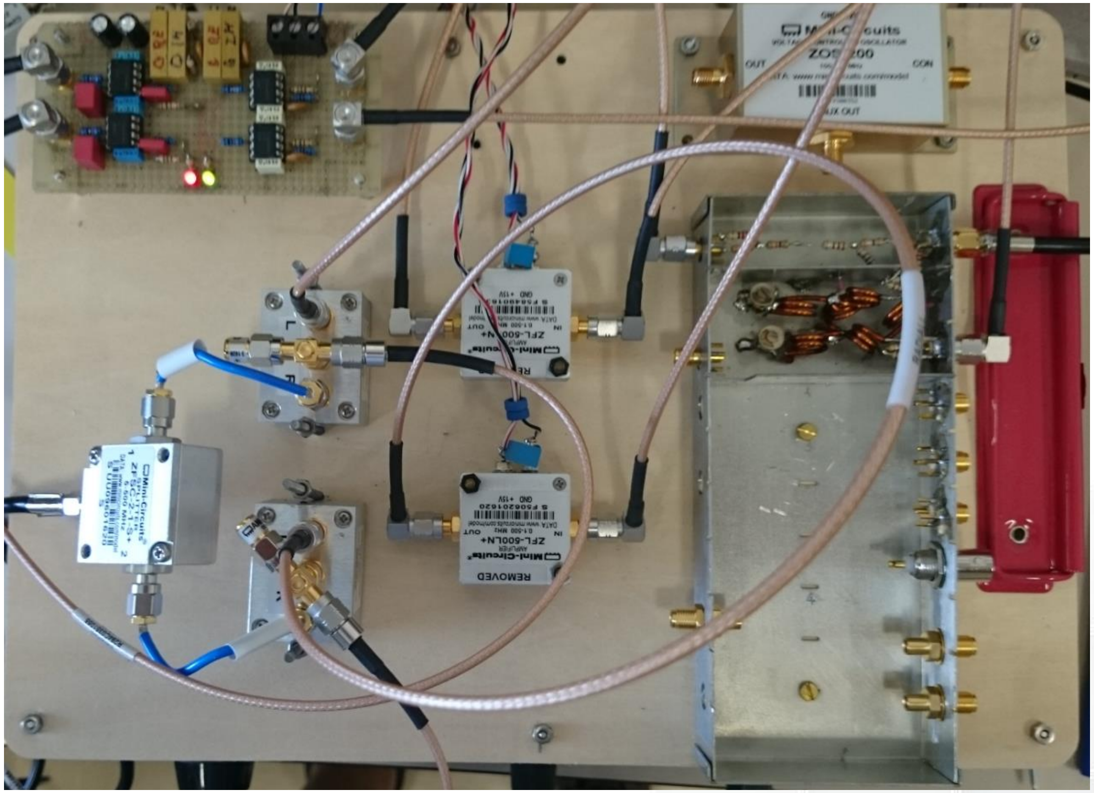




LIA Final Experimental Assembly

The first prototype has been assembled using a – literally – a breadboard. The project is still work-in-progress. The LIA can be miniaturized using components for printed circuits boards. For the moment we have excluded this solution for the impossibility to obtain a multi layer printed circuits with controlled impedance signal paths. Further, boxed and components allows quick substitutions in case of problems.





Lock-in Calibration and Performance Verification

Radio Frequency measurement are hard
RF calibration are harder
Performance verification can be very hard to reproduce

We propose calibration and performance verification procedures based on assumptions and hypothesis based on the RGB-ITR context and field experience.
It is not yet completely clear if these assumptions and hypothesis are valid in a generic context.

The **calibration procedure** is based on a 4 fixed points.

The performance verification is composed by:

- **static response**: using fixed settings, *a single parameter* (e.g.) phase is varied manually. Readings are taken "at regime" (with one seconds or greater time constant). Using different SNR, precision vs SNR tables can be produced
- **dynamic response**: using fixed settings, amplitude or phase are varied using sinusoidal stimulus signal. The recovered signal is compared with starting signal. FFT (power spectra) of the recovered signal allows SNR estimation.

Lock-in Calibration

The calibration procedure follows the set-up shown in fig.1. Set the SR844 as 30 mV f.s., 1s time constant. Set the HP REF.GEN. at 190MHz, +8.5dB output. With no signal input, set the zero trimmers (offset on the LIA filters). Use the signal attenuator gain of the simulator in order to read 20 mV(RMS) amplitude on the SR844. Turn the PM SET until the Im(LIA) is zero and the Re(LIA) is maximum. On the SR 844, set this point as zero phase (SHIFT PHASE). Adjust the ReGAIN trimmer in order to obtain the correct reading on Re(LIA).

Turn the PM-SET knob until the Re(LIA) is zero and the Im(LIA) is max. Adjust the ImGAIN trimmer in order to obtain the correct reading. Repeat the procedure on the next two max/zero points.
In summary, the four cal. points:

1. Re(LIA)=+20 mV; Im(LIA)=0 mV
2. Re(LIA)=0 mV ; Im(LIA=+20 mV
3. Re(LIA)=-20 mV; Im(LIA)=0 mV
4. Re(LIA)=0 mV; Im(LIA)=-20 mV

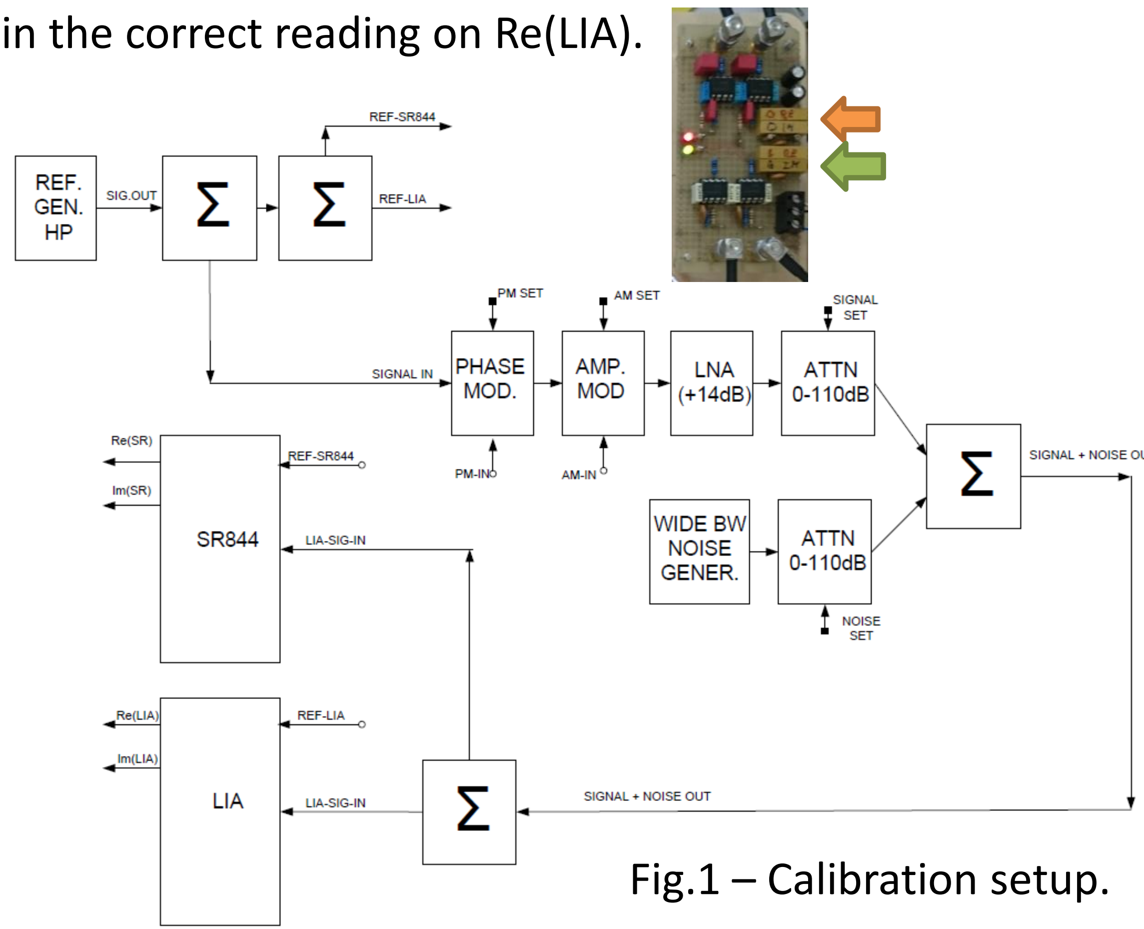


Fig.1 – Calibration setup.

Lock-in STATIC Performance Verification

The static performance verification procedure uses the same setup (fig.1). Set the SR844 as 30 mV f.s., 1s time constant. Set the HP REF.GEN. at 190MHz, +8.5dB output. Use PM-SET, AM-SET, SIGNAL-SET, NOISE-SET to establish varying operating conditions. Compare the Re(SR),Im(SR) with Re(LIA), Im(LIA).

Typical performance verification test is start with zero phase, no noise and fixed amplitude, usually 1/3 or more than the full scale.
Scan the phase 0-360°, then increase the noise contribution and repeat the test.

The resulting performances curves show how the noise degrade the static accuracy.

The very same test can be performed using different signal amplitudes, close to saturation and near the noise floor.

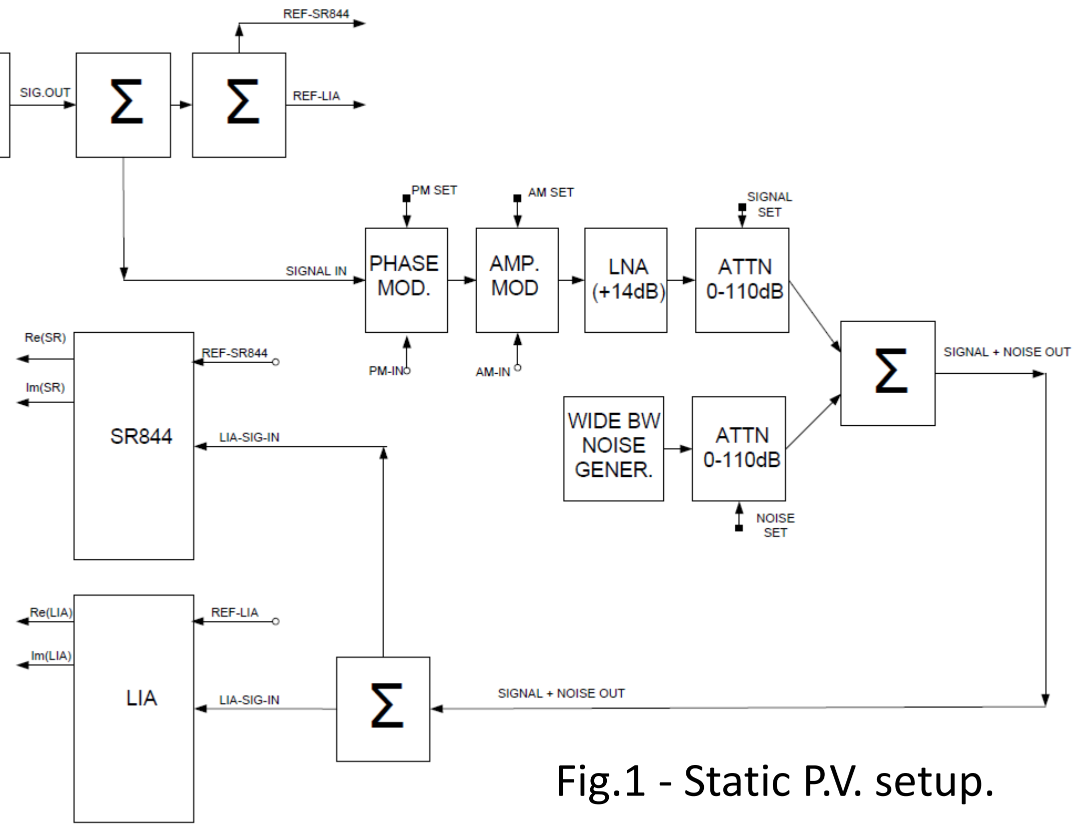
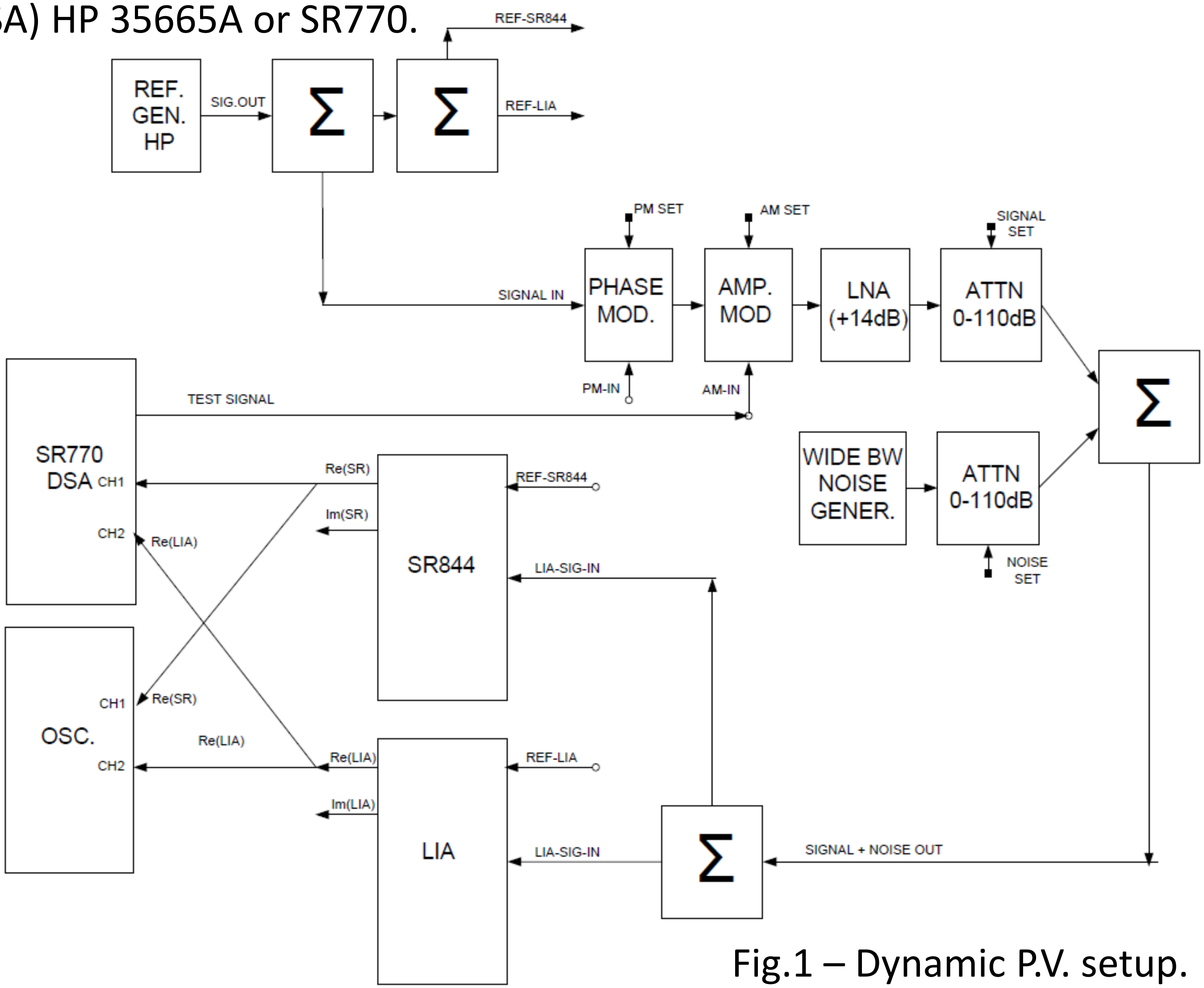


Fig.1 - Static P.V. setup.

Lock-in DYNAMIC Performance Verification

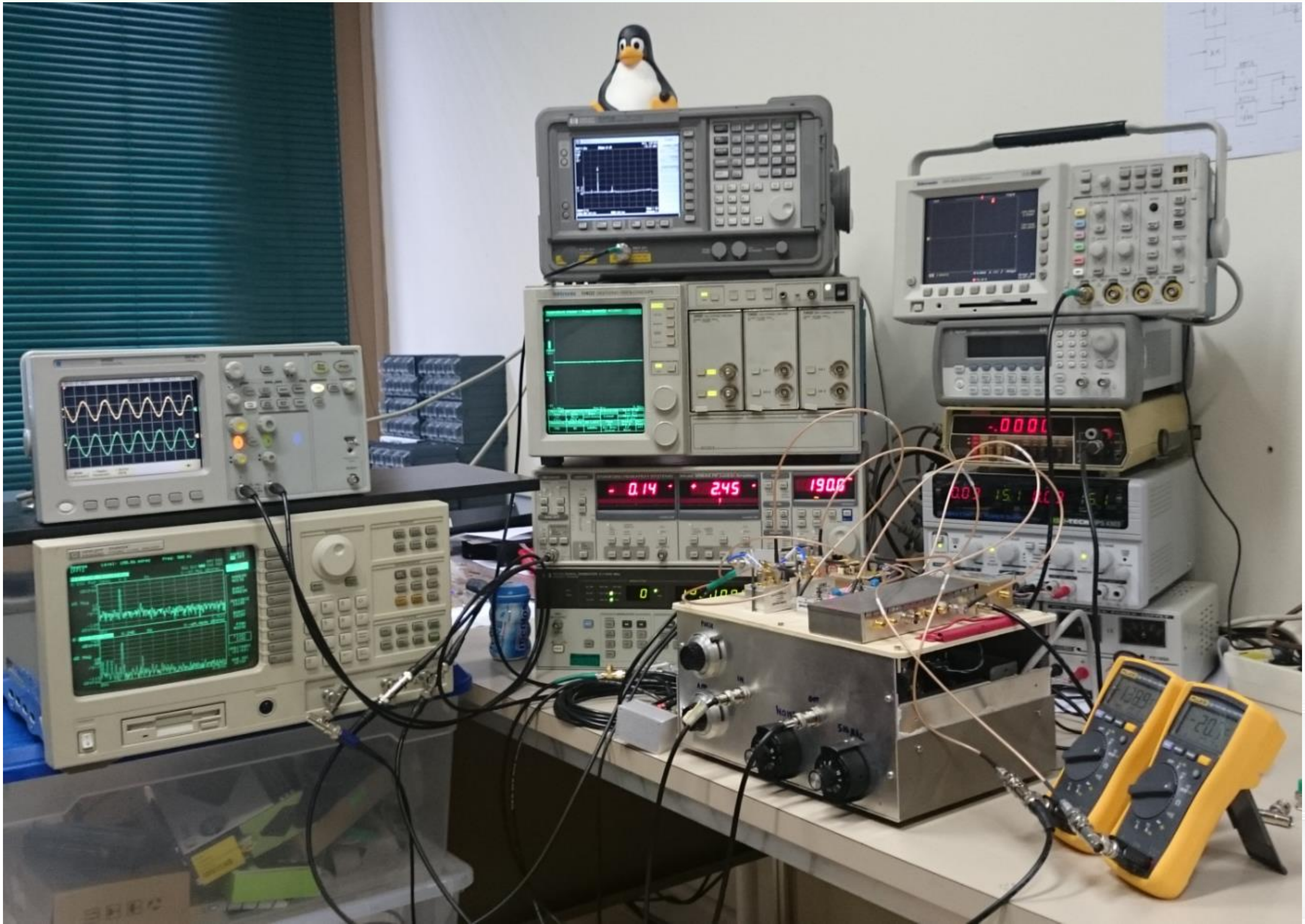
The dynamic performance verification procedure uses a more complex setup (fig. 1) that allows the estimation of the SNR of the Re (or Im) output signals. The output SNR is measured using a dynamic signal analyzer (DSA) HP 35665A or SR770.



An oscilloscope is used as monitor.

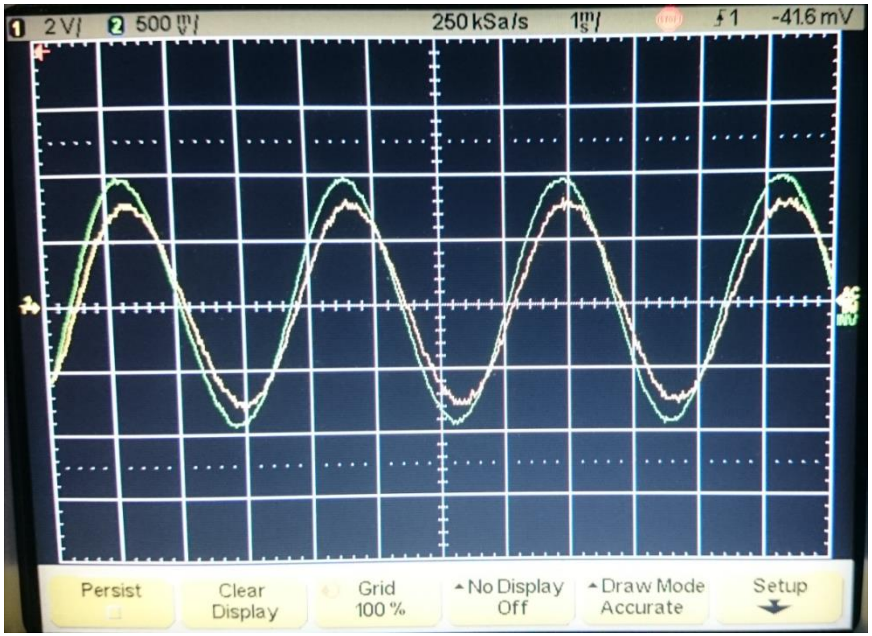
Fig.1 – Dynamic P.V. setup.

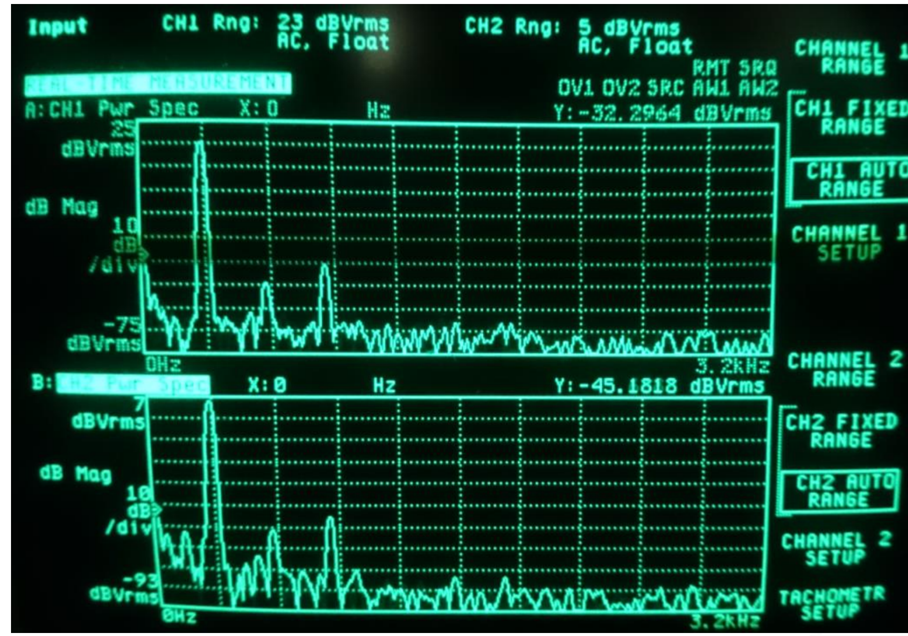
Lock-in DYNAMIC Performance Verification (April 2017)



Lock-in DYNAMIC Performance Verification

A 300 Hz signal is used to 100% modulate the 190MHz carrier signal (10 mV) simulating a rapid variation of the luminosity of the target. The phase has been tuned in order to have zero imaginary part on both LIA.





The two Re(SR,LIA) signals have different amplitude because the LIA output amplifiers are calibrated on a different scale (just for reading convenience on voltmeters used for LIA).

Signal quality (with no noise) is good (SR yellow, LIA green).

The power spectrum of the two recovered signals shown the fundamental (300 Hz), second and third harmonics. Top trace is SR, bottom trace is LIA. The noise floor (respect the fundamental) for SR is -80dB, -90dB for the LIA. The third harm. is -50 dB for SR, -55dB for LIA.

Questions without a final answer

1. Why the measured noise of the APD is TEN TIMES (approx.) the specifications?
2. Why several signal+noise measurements on RGB-ITR are not coherent between oscilloscope and spectrum analyzer? (*Some measurements must be done again*).
3. The noise floor depends from the presence of signal input (black vs white)? (*No for RED, Yes for BLU, GREEN*)
4. How to design temperature stable, tunable phase shifter for the RGB-ITR simulator?
5. The actual LIA design can be optimize further or it is already “near-the-top”?

Future developments

1. Several questions demand an answer.
2. LIA must be tested on RGB-ITR.
3. LIA should be optimized and engineered.
4. Analog lock-in works, but the future is digital/FPGA.
5. Red Pitaya is the perfect development platform for a (10 MHz max) fully digital lock-in amplifier.
6. ENEA has approved a patent application.

Conclusions

1. Simply put, develop an analog lock-in amplifiers from scratch, it is a mission for a kamikaze.
2. Probably, the development of digital/FPGA lock-in is a mission for two kamikazes.
3. Execute a wrong measure of signal + noise it is a very easy mistake to do.
4. Double or triple check everything is painstaking and very time consuming.
5. Apparently, LIA is better than Stanford (for dynamic signals), *but without a complete characterization of static errors, a definitive conclusions cannot be draw.*