Philips
RF Manual

product & design manual for
RF small signal discretes

4th edition
March 2004

APPENDIX

http://www.philips.semiconductors.com/markets/mms/products/discretes/documentation/rf_manual

Document number: 4322 252 06388
Date of release: March 2003

Content appendix:

Appendix A: 2.4GHz Generic Front-End reference design

1.1. Introduction

1.1.1. Description of the generic Front-End

This note describes the design and realization of a 2.4GHz ISM front end (Industrial-Scientific-Medical). Useful for wireless communication applications, LAN and e.g. Video/TV signal transmission. It covers power amplifier ***(PA)*** design in the Tx path, Low Noise Amplifier ***(LNA)*** design in the Rx path and RF multiplexing towards the antenna.

Though actual IC processes enable front-end integration to a certain extend, situations do exists were dedicated discrete design is required, e.g. to realize specific output power. On top of the factual design, attention is paid to interfacing the front end to existing Philips IC. More then trying to fit a target application, our intention here is to illustrate generic discrete Front end design methodology.

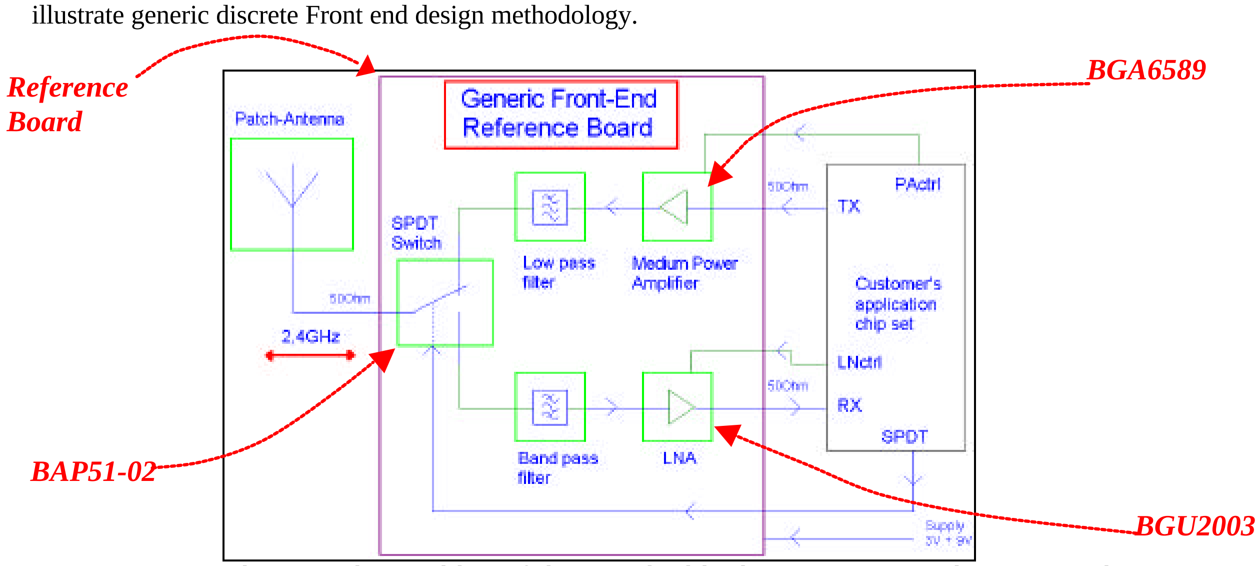


Figure1: The position of the LNA inside the 2.4GHz Generic Front-End

- ***The job of the Front-End in an application***

The board supports half duplex operation. This means the TX and RX operation are not possible at the same time. The time during TX and RX activity are so called ***time slots*** or just ***slots***. The order of the TX and RX slots is specific for the selected standard. Special handshaking activities consist of several TX and RX slots put together in to the so-called ***time-frame*** or just ***frame***. The ***user points*** / ***access points*** linked in this wireless application must follow the same functionality of slots, same order of frames and timing procedure ***(synchronization)***. These kind of issues must be under the control of specific rules (standard) normally defined by ***Institutes*** or ***Organization*** like ETSI, IEEE, NIST, FCC, CEPT, and so on.

- ***How does the Front-End work?***

Under the control (SPDT-PIN) of customer's chip set, the Front-End ***SPDT*** (**S**ingle **P**ole **D**ouble **T**hrough) Switch based on the ***PIN Diode BAP51-02*** closes the path between the antenna and the Medium Power Amplifier in the TX time slot. The PA can be switched on/off by the PAVcc-PIN. The output power signals can be radiated from the antenna away into the Ether/Space. The ***Ether*** is the natural environment medium around being used by the wireless RF traveling waves from one access point to the other one. Because the TX signals are amplified by the ***Medium Power Amplifier BGA6589,*** more powerful signals can be transmitted and reach further distances. The signal receiving occurs during the RX time slot. For this operation mode, the antenna is switched away from the **PA** (power amplifier) and connected to the LNA input under the control of the SPDT-PIN. The LNA can be switched on/off by the LNVcc-PIN. System analysis on a receiver noise performance can show that a ***low noise amplifier (LNA) BGU2003*** does improve the receiver's sensitivity by reduction of the effective RX system noise figure ***(NF)***. That's possible by installing moderate gain with very low noise in the front of the noisy input receiver IC by the use of the LNA. The effect is the receiver's ability of properly receive signals from access points at much further distances. This effect can be shown by the mathematical relationship shown below:

With the general Noise Figure **(NF)** definition: $NF = 10 \cdot \log(F) = 10\log\left(\frac{Pout_{Noise}}{Pin_{Noise}}\right)$. All the time, the amount of the noise ratio F will be larger than one (*F>1 or NF>0dB*) for operating at temperature larger than zero degree Kelvin.

The overall System Noise Ratio of the cascade LNA + RX chip results in: $F_{SYST} = F_{LNA} + \frac{F_{RX} - 1}{Gain_{LNA}}$ The F_{SYST} illustrates that the overall system noise ratio (LNA+RX chip set) is at least the F_{LNA}. There is the addition of a second amount of noise caused by the ICs RX channel. But this amount is reduced by the LNA gain $Gain_{LNA}$. Use of moderate LNA does reduce the noise ratio part of the receiver chip set. In this kind of relationship the LNA's noise ratio F_{LNA} is dominant.

Example-1:

- *Issue:* Customer's receiver chip-set with a NF=9dB; LNA with Power Gain=13dB and NF=1.3dB
- *Question:* What's the amount of the system receiver's noise figure?
- *Calculation:*

$$Gain_{LNA} = 10^{\frac{13dB}{10}} = \underline{20}$$

$$F_{RX} = 10^{\frac{9dB}{10}} = \underline{7.943}$$

$$F_{LNA} = 10^{\frac{1.3dB}{10}} = \underline{1.349}$$

LNA-noise part

Shrank RX chip-set's adding noise part

$$F_{SYST} = F_{LNA} + \frac{F_{RX} - 1}{Gain_{LNA}} = 1.349 + \frac{7.943 - 1}{20} = 1.349 + 0.347$$

$$\underline{F_{SYST} = 1.696}$$

$$NF_{SYST} = 10\log(F_{SYST}) = 10\log(1.696)$$

$$\underline{\underline{NF_{SYST} = 2.3dB}}$$

- *Answer:* In this example the use of the LNA in front of the receiver chip-set does improve the overall receiver system Noise Figure to NF=2.3dB. The equations show that the first device in a cascade of objects has the most effect on the overall noise figure. In reality the first part of a receiver is the antenna. Its quality is very important.**Example-2:**

Philips Medium Power MMICs portfolio offer the following listed insertion power gain $|S_{21}|^2$ performances:

BGA6289 ➔ 12dB
BGA6589 ➔ 15dB

- *Question:*

What is the expected approximated increase of distance using this Philips' MMICs negating the attenuation of the Ether from an antenna with 3D homogenous round around field radiation in front of the chip-set?

- *Evaluation:*

3D homogenous round around radiation power is general done by an ideal spherical dot. The following law describes theoretical the power-density of damped traveling waves, radiated by the reference-isotropic antenna in a certain distance:

$$P_{E(r)} = P_S \cdot A_E \cdot \frac{1}{4p \cdot r^2} \cdot e^{-c \cdot r}$$

$P_{E(r)}$ = Receiver power in the distance "r" to the transmitter's isotropic antenna
r = Distance receiver-transmitter
P_S = Transmitter power
χ = Atmospheric attenuation exponent
A_E = Receiver antenna surface

This kind of general Physic's law is used for all kinds of spherical wave and energy radiation topics like in optics, acoustics, thermal, electromagnetic and so on. The job of the electromagnetic wave radiating antenna is the power matching of the cable impedance (50Ω, 75Ω,...) to the space's impedance with the (ideal) electromagnetic far field impedance of 120πΩ.
The received normalized power/unit area P_r at the receiver transmitted from a transmitter with the power P_t in the distance d and neglecting of atmospheric attenuation (χ=0) is calculated by: $P_{RX} = \frac{P_{TX}}{4p \cdot d^2}$

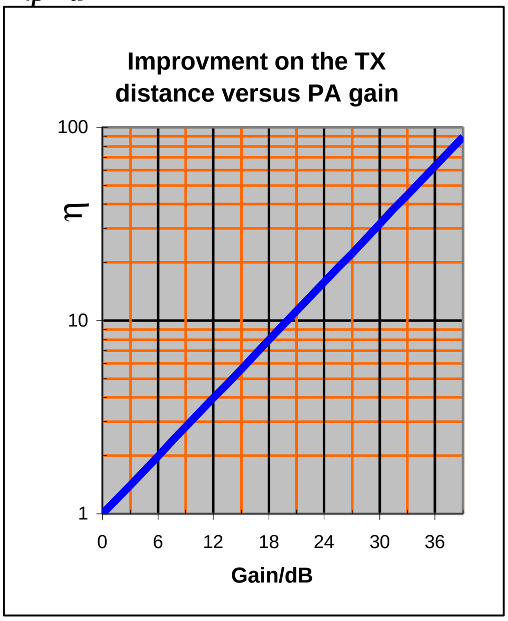
TX-RX-distance: ➔ $d = \sqrt{\frac{P_t}{4p \cdot P_r}}$ without PA: $d_1 = \sqrt{\frac{P_{t1}}{4p \cdot P_r}}$

Expanded distance by the PA for same received RX power: $d_2 = \sqrt{\frac{P_{t2}}{4p \cdot P_r}}$

$$h = \frac{d_2}{d_1} = \frac{\sqrt{\frac{P_{t2}}{4p \cdot P_r}}}{\sqrt{\frac{P_{t1}}{4p \cdot P_r}}} = \sqrt{|S21|^2} \qquad ➔ \qquad h = \sqrt{|S21|^2}$$

BGA6289 gain factor: $10^{\frac{12dB}{10}} = 15.85$ BGA6589 gain factor: $10^{\frac{15dB}{10}} = 31.62$

$h_{BGA6289} = \sqrt{15.85} = 3.98$ $\qquad h_{BGA6589} = \sqrt{31.62} = 5.62$



- *Answer:*

Use of BGA6289 can theoretical increase the transmitter operation area by the factor of 4. The BGA6589 can increase the operation area by 5.6 assuming no compression of the amplifiers and an isotropic antenna radiator. In reality we have to take into account the amplifier input/output matching circuits adding or removing of gain to device's insertion power gain, the frequency depending attenuation of the Ether and the gain of the receiver and transmitter antenna.

1.1.2 Applications for the Reference Board

Some application ideas for the use of the Generic Front-End Reference Board

- 2.4GHz WLAN
- Wireless video, TV and remote control signal transmission
- PC to PC data connection
- PC headsets
- PC wireless mouse, key board, and printer
- Palm to PC, Keyboard, Printer connectivity
- Supervision TV camera signal transmission
- Wireless loudspeakers
- Robotics
- Short range underground walky-talky
- Short range snow and stone avalanche person detector
- Key less entry
- Identification
- Tire pressure systems
- Garage door opener
- Remote control for alarm-systems
- Intelligent kitchen (cooking place, Microwave cooker and washing machine operator reminder)
- Bluetooth
- DSSS 2.4GHz WLAN (IEEE802.11b)
- OFDM
- 2.4GHz WLAN (IEEE802.11g)
- Access Points
- PCMCIA
- PC Cards
- 2.4GHz Cordless telephones
- Wireless pencil as an input for Palms and PCs
- Wireless hand scanner for a Palm
- Identification for starting the car engine
- Wireless reading of gas counters
- Wireless control of soft-drink /cigarette/snag - SB machine
- Communication between bus/taxi and the stop lights
- Panel for ware house stock counting
- Printers
- Mobiles
- Wireless LCD Display
- Remote control
- Cordless Mouse
- Automotive, Consumer, Communication

Please note:
The used MMICs and PIN diodes can be used in other frequency ranges e.g. 300MHz to 3GHz for applications like communication, networking and ISM too.

1.1.2. The Reference Board together with Philips ICs

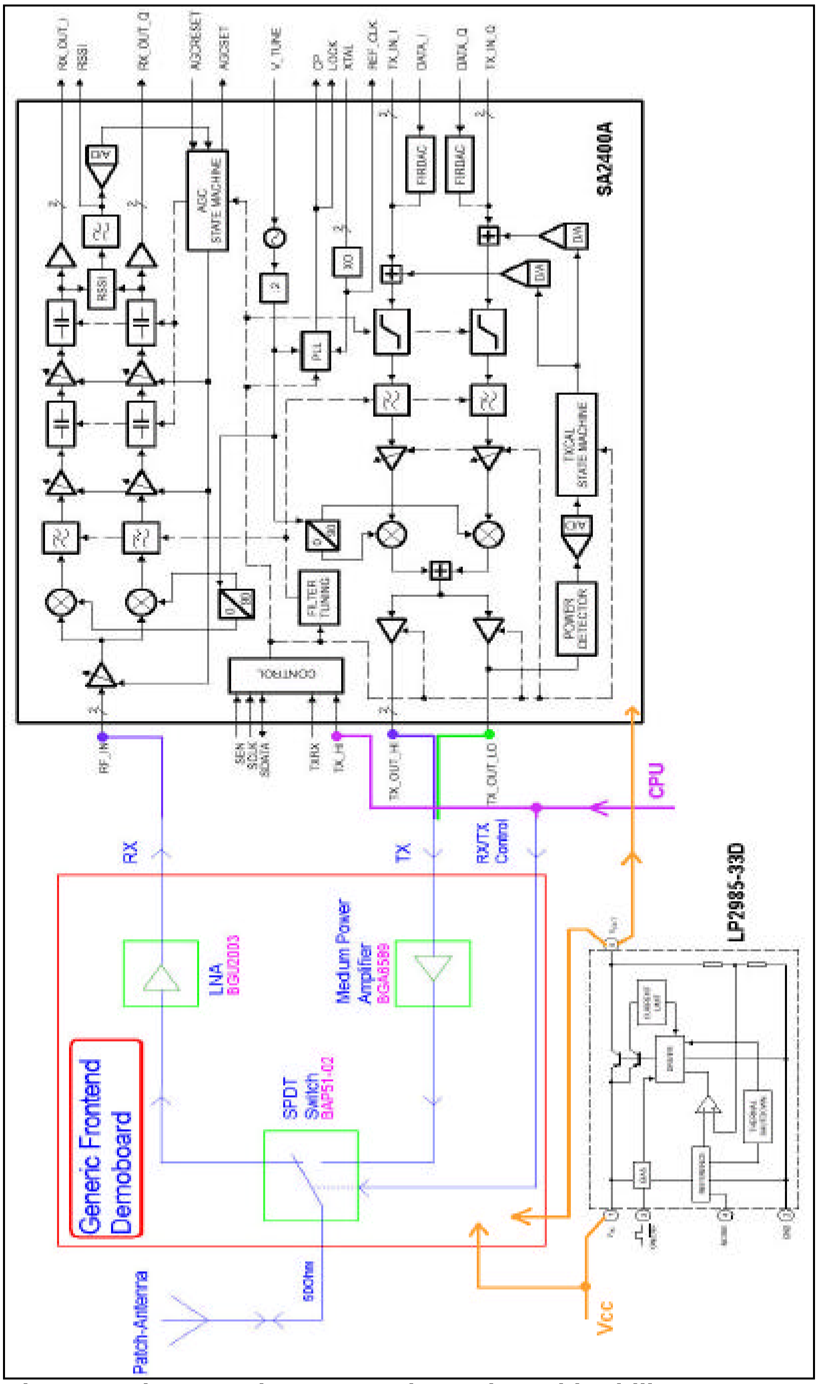


Figure 2: The Generic Front-End together with Philips' SA2400A for 2.45GHz ISM band

Illustrated is a principle idea how the 2.4GHz Generic Front-End Reference Board can work together with a transceiver for improved performances.

Up and down direct conversion I/Q transmitter for 2.4GHz with TX output power up to +20dBm and RX low noise. Digital control of all functions.

Main devices from Philips Semiconductors:

- BGU2003
- BGA6589
- BAP51-02
- SA2400A
- LP2985-33D

1.1.3. Selection of Applications in the 2.4GHz environment

Application	Standardization name/ issue	Start frequency	Stop Frequency	Centre frequency	Bandwidth-MHz/ Channel Spacing-MHz
Bluetooth; 1Mbps	IEEE802.15.1	NUS/EU=2402MHz (All)=2402MHz	NUS/EU=2480MHz (All)=2495MHz	2442.5MHz	NUS/EU=78/1MHz (All)=93/1MHz
WiMedia , (802.15.3a@3.1-10.6GHz)	IEEE802.15.3 (camera, video)	2.4GHz	2.49GHz	2.45GHz	
ZigBee; 1000kbps@2450MHz Other Frequency(868; 915)MHz	IEEE802.15.4	US=2402MHz EU=2412MHz	US=2480MHz EU=2472MHz	2441MHz	US=83/4MHz EU=60/4
DECT@ISM	ETSI	2400 MHz	2483MHz	2441.5MHz	83/
IMT-2000 =3G; acc., ITU, CEPT, ERC ERC/DEC/(97)07; ERC/DEC/(99)25 (=UMTS, CDMA2000, UWC-136, UTRA-FDD, UTRA-TDD)	FDD Uplink (D)	≈1920	≈1980	Exact Frequency range depending on country & system supplier	(TDD, FDD; WCDMA, TD-CDMA); paired 2x60MHz (D) non paired 25MHz (D)
	FDD Downlink (D)	≈2110	≈2170		
	TDD (D)	≈1900	≈2024		
USA - ISM		2400MHz	2483.5MHz	2441.75MHz	83.5/
Wireless LAN; Ethernet; (5.2; 5.7)GHz	IEEE802.11; (a, b, …)	2400MHz	2483MHz	2441.5MHz	83/FHSS=1MHz; DSSS=25MHz
Wi-Fi; 11-54Mbs; (4.9-5.9)GHz	IEEE802.11b; (g, a)	2400MHz	2483MHz	2441.5MHz	
RFID	ECC/SE24	2446MHz	2454MHz	2.45GHz	
Wireless LAN; 11Mbps	IEEE802.11b	2412MHz	2462MHz	2437MHz	56/
Wireless LAN; 54Mbps	IEEE802.11g				
WPLAN	NIST	2400MHz			
HomeRF; SWAP/CA, 0.8-1.6Mbps		NUS/EU=2402MHz (All)=2402	NUS/EU=2480MHz (All)=2495		78/1MHz, 3.5MHz 93/1MHz, 3.5MHz
Fixed Mobile; Amateur Satellite; ISM, SRD, RLAN, RFID	ERC, CEPT Band Plan	2400MHz	2450MHz	2425MHz	50/
Fixed RF transmission	acc. CEPT Austria regulation	2400MHz	2450MHz	2425MHz	50/
MOBIL RF; SRD	acc. CEPT Austria regulation	2400MHz	2450MHz	2425MHz	50/
Amateur Radio	FCC	2390MHz	2450MHz		60/
UoSAT-OSCAR 11, Telemetry	Amateur Radio Satellite UO-11			2401.5MHz	
AMSAT-OSCAR 16	Amateur Radio Satellite AO-16			2401.1428MHz	
DOVE-OSCAR 17	Amateur Radio Satellite DO-17			2401.2205MHz	
Globalstar, (Mobile Downlink)	Loral, Qualcomm	2483.5MHz	2500MHz	S-Band	
Ellipso, (Mobile Downlink)	Satellite; Supplier Ellipsat	2483.5MHz	2500MHz		
Aries, (Mobile Downlink) (now Globalstar?)	Satellite; Supplier Constellation	2483.5MHz	2500MHz		
Odyssey, (Mobile Downlink)	Satellite; Supplier TRW	2483.5MHz	2500MHz		
Orbcomm Satellite (LEO) eg. GPSS-GSM	Satellite			2250,5MHz	
Ariane 4 and Ariane 5 (ESA, Arianespace)	tracking data link for rocket			2206MHz	
Atlas Centaur eg. carrier for Intelsat IVA F4	tracking data link for rocket			2210,5MHz	
J.S. Marshall Radar Observatory	700KW Klystron TX			S-Band	
Raytheon ASR-10SS Mk2 Series S-Band Solid-State Primary Surveillance Radar	US FAA/DoD ASR-11 used in U.S. DASR program	2700	2900	S-Band Radar ≈2400MHz	
Phase 3D; Amateur Radio Satellite; 146MHz, 436MHz, 2400MHz	AMSAT; 250Wpep TX			S-Band	2.4KHz, SSB
Apollo 14-17; NASA space mission	transponder experiments			S-Band	
ISS; (internal Intercom System of the ISS station)	Space			2.4GHz	
MSS Downlink	UMTS	2170	2200		

Abbreviations: European Radio communication Committee **(ERC)** within the European Conference of Postal and Telecommunication Administration **(CEPT)**

NIST	=	National Institute of Standards and Technology	***RFID***	=	Radio Frequency Identification
WPLAN	=	Wireless Personal Area Networks	***OSCAR***	=	Orbit Satellite Carry Amateur Radio
WLAN	=	Wireless Local Area Networks	***FHSS***	=	Frequency Hopping Spread Spectrum
ISM	=	Industrial Scientific Medical	***DSSS***	=	Direct Sequence Spread Spectrum
LAN	=	Local Area Network	***DECT***	=	Digital Enhanced Cordless Telecommunications
IEEE	=	Institute of Electrical and Electronic Engineers	***NUS***	=	North America
SRD	=	Short Range Device	***EU***	=	Europe
RLAN	=	Radio Local Area Network	***ITU***	=	International Telecommunications Union
ISS	=	International Space Station	***ITU-R***	=	ITU Radio communication sector
IMT	=	International mobile Telecommunications at 2000MHz	***(D)***	=	Germany
MSS	=	Mobile Satellite Service	***TDD***	=	Time Division Multiplex
W-CDMA	=	Wideband-CDMA	***FDD***	=	Frequency Division Multiplex
GMSK	=	Gaussian Minimum Shift Keying	***TDMA***	=	Time Division Multiplex Access
UMTS	=	Universal Mobile Telecommunication System	***CDMA***	=	Code Division Multiplex Access
UWC	=	Universal Wireless Communication	***2G***	=	Mobile Systems GSM, DCS
MSS Downlink	=	Mobile Satellite Service of UMTS	***3G***	=	IMT-2000

1.2. Summary

1.2.1. Block Diagram

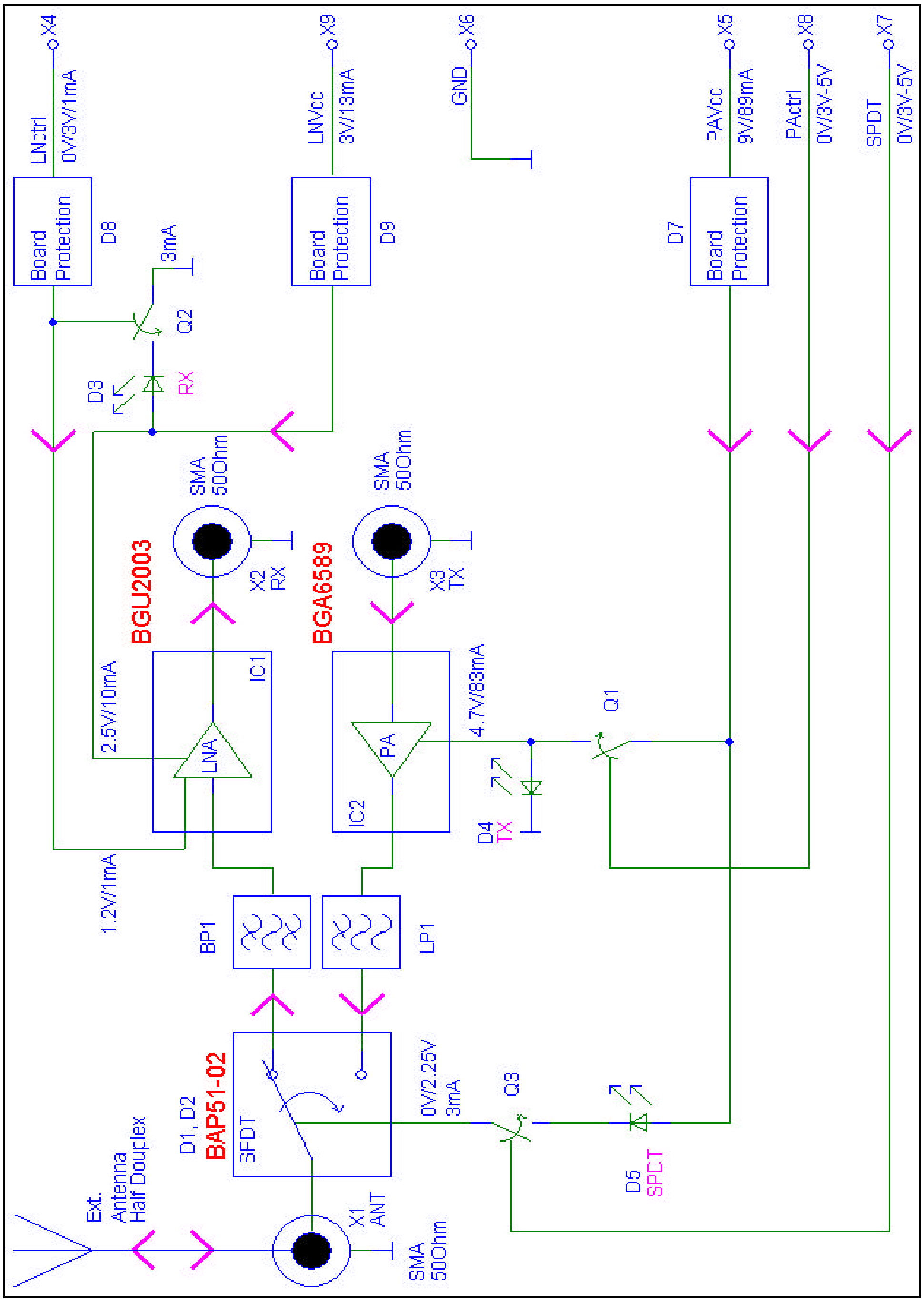


Figure 3: Block Diagram of the Reference Board

1.2.2. Schematic

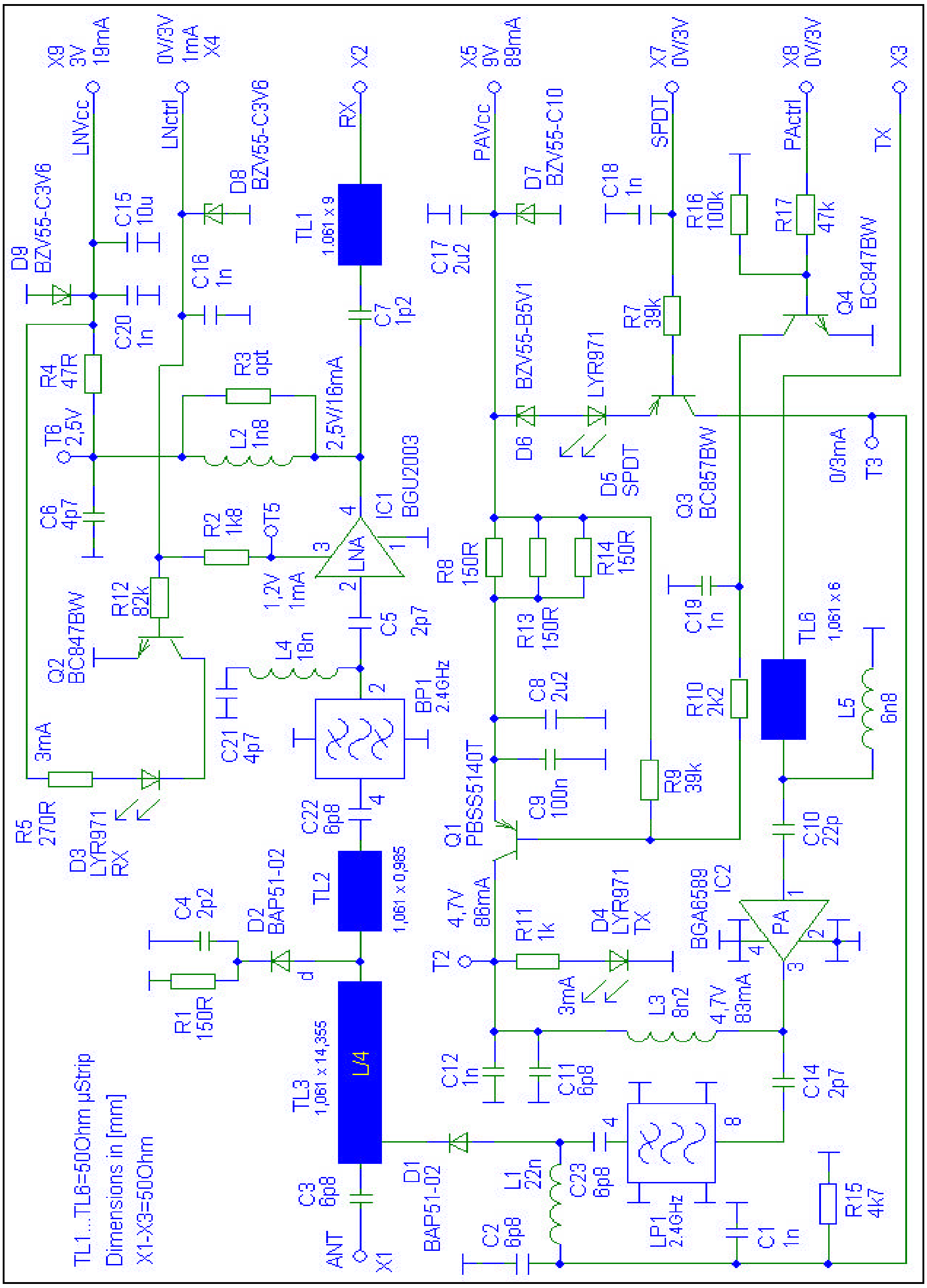


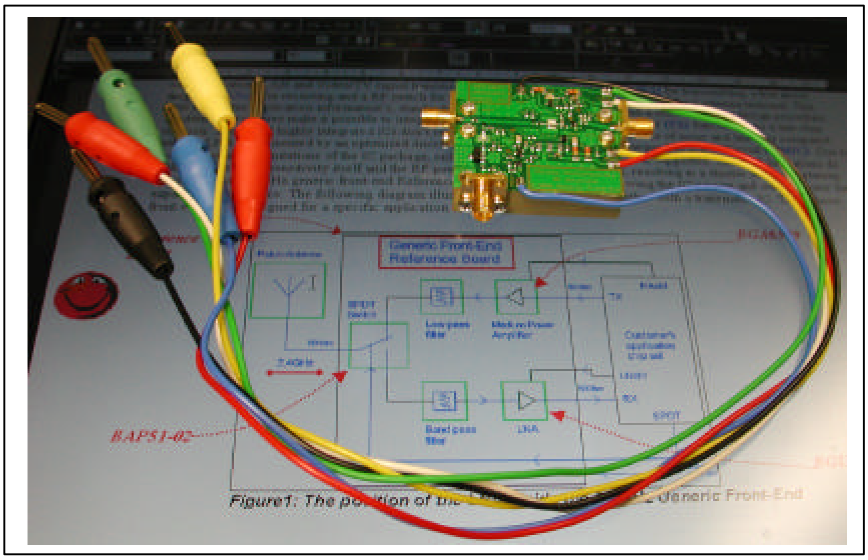
Figure 4: Schematic of the Reference Board

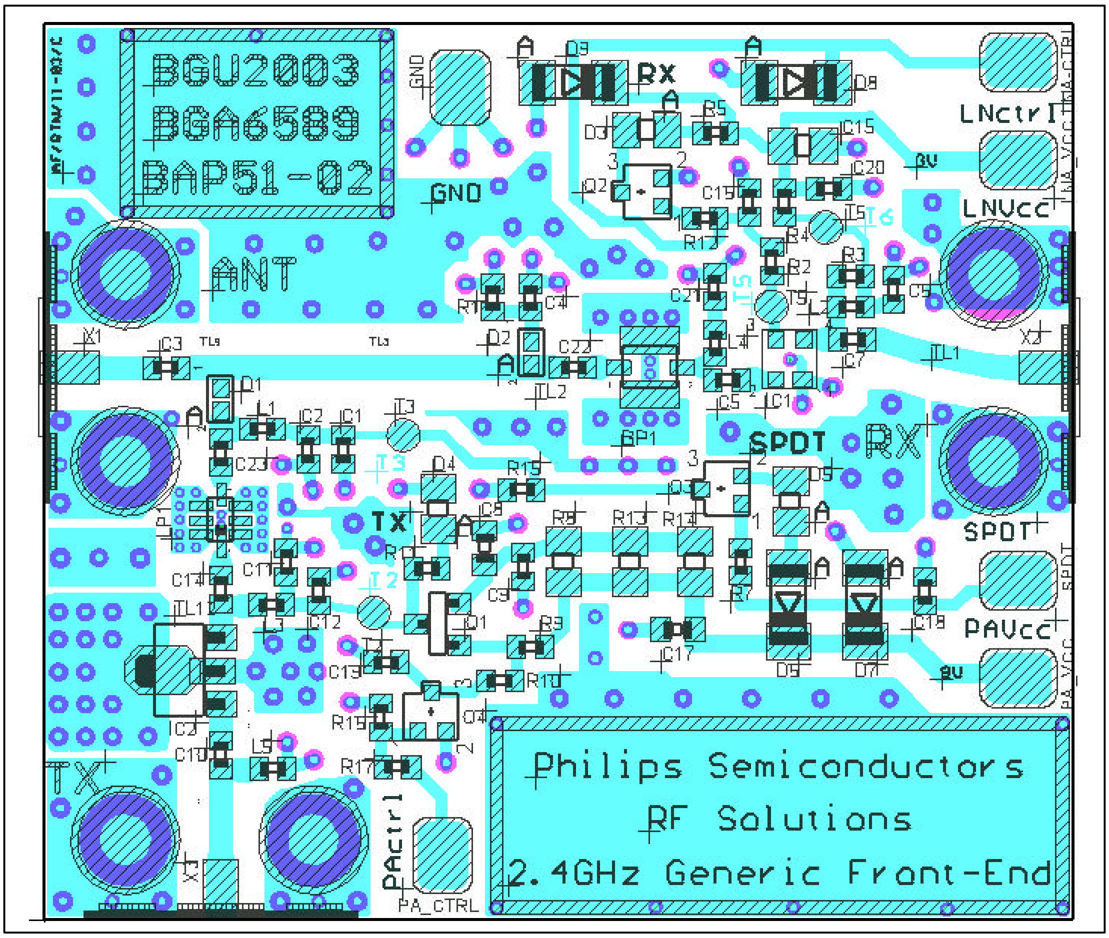
1.2.3. Part List

Part Number	Value	Size	Function / Short explanation	Manufacturer	Order Code	Order source
IC1	BGU2003	SOT363	LNA-MMIC	Philips Semiconductors	BGU2003	PHL
IC2	BGA6589	SOT89	TX-PA-MMIC	Philips Semiconductors	BGA6589	PHL
Q1	PBSS5140T	SOT23	TX PA-standby control	Philips Semiconductors	PBSS5140T	PHL
Q2	BC847BW	SOT323	Drive of D3	Philips Semiconductors	BC847BW	PHL
Q3	BC857BW	SOT323	SPDT switching	Philips Semiconductors	BC857BW	PHL
Q4	BC847BW	SOT323	PA logic level compatibility	Philips Semiconductors	BC847BW	PHL
D1	BAP51-02	SOD523	SPDT-TX; series part of the PIN diode switch	Philips Semiconductors	BAP51-02	PHL
D2	BAP51-02	SOD523	SPDT-RX; shunt part of the PIN diode switch	Philips Semiconductors	BAP51-02	PHL
D3	LYR971	0805	LED, yellow, RX and bias current control of IC1	OSRAM	67S5126	Bürklin
D4	LYR971	0805	LED, yellow; TX	OSRAM	67S5126	Bürklin
D5	LYR971	0805	LED, yellow; SPDT; voltage level shifter	OSRAM	67S5126	Bürklin
D6	BZV55-B5V1	SOD80C	Level shifting for being 3V/5V tolerant	Philips Semiconductors	BZV55-B5V1	PHL
D7	BZV55-C10	SOD80C	Board DC polarity & over voltage protection	Philips Semiconductors	BZV55-C10	PHL
D8	BZV55-C3V6	SOD80C	Board DC polarity & over voltage protection	Philips Semiconductors	BZV55-C3V6	PHL
D9	BZV55-C3V6	SOD80C	Board DC polarity & over voltage protection	Philips Semiconductors	BZV55-C3V6	PHL
R1	150Ω	0402	SPDT bias	Yageo RC0402 Vitrohm512	26E558	Bürklin
R2	1k8	0402	LNA MMIC current CTRL	Yageo RC0402 Vitrohm512	26E584	Bürklin
R3	optional	0402	L2 resonance damping; optional	---	optional	
R4	47Ω	0402	LNA MMIC collector bias	Yageo RC0402 Vitrohm512	26E546	Bürklin
R5	270Ω	0402	RX LED current adj.	Yageo RC0402 Vitrohm512	26E564	Bürklin
R7	39k	0402	Q3 bias SPDT	Yageo RC0402 Vitrohm512	26E616	Bürklin
R8	150Ω	0805	PA-MMIC collector current adjust and temperature compensation	Yageo RC0805 Vitrohm503	11E156	Bürklin
R9	39k	0402	Helps switch off of Q1	Yageo RC0402 Vitrohm512	26E616	Bürklin
R10	2k2	0402	Q1 bias PActrl	Yageo RC0402 Vitrohm512	26E586	Bürklin
R11	1kΩ	0402	LED current adjust; TX-PA	Yageo RC0402 Vitrohm512	26E578	Bürklin
R12	82k	0402	Q2 drive	Yageo RC0402 Vitrohm512	26E624	Bürklin
R13	150Ω	0805	PA-MMIC collector current adjust	Yageo RC0805 Vitrohm503	11E156	Bürklin
R14	150Ω	0805	PA-MMIC collector current adjust	Yageo RC0805 Vitrohm503	11E156	Bürklin
R15	4k7	0402	Improvement of SPDT-Off	Yageo RC0402 Vitrohm512	26E594	Bürklin
R16	100k	0402	PActrl; logic level conversion	Yageo RC0402 Vitrohm512	26E626	Bürklin
R17	47k	0402	PActrl; logic level conversion	Yageo RC0402 Vitrohm512	26E618	Bürklin
L1	22nH	0402	SPDT RF blocking for biasing	Würth Elektronik, WE-MK	744 784 22	WE
L2	1n8	0402	LNA output matching	Würth Elektronik, WE-MK	744 784 018	WE
L3	8n2	0402	PAout Matching	Würth Elektronik, WE-MK	744 784 082	WE
L4	18nH	0402	LNA input match	Würth Elektronik, WE-MK	744 784 18	WE
L5	6n8	0402	PA input matching	Würth Elektronik, WE-MK	744 784 068	WE
C1	1nF	0402	medium RF short for SPDT bias	Murata, X7R	GRP155 R71H 102 KA01E	Murata
C2	6p8	0402	medium RF short for SPDT bias	Murata, C0G	GRP1555 C1H 6R8 DZ01E	Murata
C3	6p8	0402	Antenna DC decoupling	Murata, C0G	GRP1555 C1H 6R8 DZ01E	Murata
C4	2p2	0402	RF short SPDT shunt PIN	Murata, C0G	GRP1555 C1H 2R2 CZ01E	Murata
C5	2p7	0402	DC decoupling LNA input + match	Murata, C0G	GRP1555 C1H 2R7 CZ01E	Murata
C6	4p7	0402	RF short output match	Murata, C0G	GRP1555 C1H 4R7 CZ01E	Murata
C7	1p2	0402	LNA output matching	Murata, C0G	GRP1555 C1H 1R2 CZ0E	Murata
C8	2u2/10V	0603	Removes the line ripple together with R8-R14 from PA supply rail	Murata, X5R	GRM188 R61A 225 KE19D	Murata
C9	100nF/16V	0402	Ripple rejection PA	Murata, Y5V	GRM155 F51C 104 ZA01D	Murata
C10	22pF	0402	DC decoupling PA input	Murata, C0G	GRP1555 C1H 220 JZ01E	Murata
C11	6p8	0402	RF short-bias PA	Murata, C0G	GRP1555 C1H 6R8 DZ01E	Murata
C12	1nF	0402	PA, Supply RF short	Murata, X7R	GRP155 R71H 102 KA01E	Murata

Part Number	Value	Size	Function / short explanation	Manufacturer	Order Code	Order source
C14	2p7	0402	TX-PAout DC decoupling + matching	Murata, C0G	GRP1555 C1H 2R7 CZ01E	Murata
C15	10u/6.3V	0805	dc rail LNVcc	Murata, X5R	GRM21 BR60J 106 KE19B	Murata
C16	1nF	0402	dc noise LNctrl	Murata, X7R	GRP155 R71H 102 KA01E	Murata
C17	2u2/10V	0603	PA dc rail	Murata, X5R	GRM188 R61A 225 KE34B	Murata
C18	1nF	0402	dc noise SPDT control	Murata, X7R	GRP155 R71H 102 KA01E	Murata
C19	1nF	0402	dc noise PActrl	Murata, X7R	GRP155 R71H 102 KA01E	Murata
C20	1nF	0402	dc noise LNVcc	Murata, X7R	GRP155 R71H 102 KA01E	Murata
C21	4p7	0402	RF short for optional LNA input match	Murata, C0G	GRP1555 C1H 4R7 CZ01E	Murata
C22	6p8	0402	dc removal of RX-BP filter and matching	Murata, C0G	GRP1555 C1H 6R8 DZ01E	Murata
C23	6p8	0402	dc removal of TX-LP filter and matching	Murata, C0G	GRP1555 C1H 6R8 DZ01E	Murata
BP1	fo=2.4GHz	1008	RX band pass input filtering	Würth Elektronik	748 351 024	WE
LP1	fc=2.4GHz	0805	TX low pass spurious filtering	Würth Elektronik	748 125 024	WE
X1	SMA, female µStrip tab pin	12.7mm flange 1.3mm tab	Antenna connector, SMA, panel launcher, female, bulkhead receptacle with flange, PTFE, CuBe, CuNiAu	Telegärtner	J01 151 A08 51	Telegärtner
X2	SMA, female µStrip tab pin	12.7mm flange 1.3mm tab	RX-Out connector, SMA, panel launcher, female, bulkhead receptacle with flange, PTFE, CuBe, CuNiAu	Telegärtner	J01 151 A08 51	Telegärtner
X3	SMA, female µStrip tab pin	12.7mm flange 1.3mm tab	TX-IN connector, SMA, panel launcher, female, bulkhead receptacle with flange, PTFE, CuBe, CuNiAu	Telegärtner	J01 151 A08 51	Telegärtner
X4	BÜLA30K	green	LNctrl, BÜLA30K, Multiple spring wire plugs, Solder terminal	Hirschmann	15F260	Bürklin
X5	BÜLA30K	red	PAVcc, BÜLA30K, Multiple spring wire plugs, Solder terminal	Hirschmann	15F240	Bürklin
X6	BÜLA30K	black	GND, BÜLA30K, Multiple spring wire plugs, Solder terminal	Hirschmann	15F230	Bürklin
X7	BÜLA30K	yellow	SPDT, BÜLA30K, Multiple spring wire plugs, Solder terminal	Hirschmann	15F250	Bürklin
X8	BÜLA30K	blue	PActrl, BÜLA30K, Multiple spring wire plugs, Solder terminal	Hirschmann	15F270	Bürklin
X9	BÜLA30K	red	LNVcc, BÜLA30K, Multiple spring wire plugs, Solder terminal	Hirschmann	15F240	Bürklin
Y1	blue { PActrl }	40cm, 0.5qmm	Insulated stranded hook-up PVC wire, LiYv, blue, CuSn	VDE0812/9.72	92F566	Bürklin
Y2	red { PAVcc }	40cm, 0.5qmm,	Insulated stranded hook-up PVC wire, LiYv, red, CuSn	VDE0812/9.72	92F565	Bürklin
Y3	green { LNctrl }	40cm, 0.5qmm,	Insulated stranded hook-up PVC wire, LiYv, green, CuSn	VDE0812/9.72	92F567	Bürklin
Y4	black { GND }	40cm, 0.5qmm	Insulated stranded hook-up PVC wire, LiYv, black, CuSn	VDE0812/9.72	92F564	Bürklin
Y5	yellow { SPDT }	40cm, 0.5qmm,	Insulated stranded hook-up PVC wire, LiYv, yellow, CuSn	VDE0812/9.72	92F568	Bürklin
Y6	white { LNVcc }	40cm, 0.5qmm,	Insulated stranded hook-up PVC wire, LiYv, white, CuSn	VDE0812/9.72	92F569	Bürklin
Z1 - Z6	M2	M2 x 3mm	Screw for PCB mounting	Paul-Korth GmbH	NIRO A2 DIN7985-H	Paul-Korth
Z7 - Z12	M2,5	M2,5 x 4mm	Screw for SMA launcher mounting	Paul-Korth GmbH	NIRO A2 DIN7985-H	Paul-Korth
W1	FR4 compatible	47,5mm X 41,5mm	Epoxy 560µm; Cu=17.5µm; Ni=5µm; Au=0.3µm two layer double side	www.isola.de www.haefele-leiterplatten.de	DURAVER®-E-Cu, Qualität 104 MLB-DE 104 ML/2	Häfele Leiterplat-tentechnik
W2	Aluminum metal finished yellow Aludine	47,5mm X 41,5mm X 10mm	Base metal caring the pcb and SMA connectors	---	---	---

1.2.4. The PCB




BGU2003
BGA6589
BAP51-02
ANT
GND
RX
TX
SPDT
LNCtrl
LNVcc
PAVcc
PActrl
PA_CTRL
Philips Semiconductors
RF Solutions
2.4GHz Generic Front-End

1.2.5. Functional description

1.2.5.1. Principle of operation

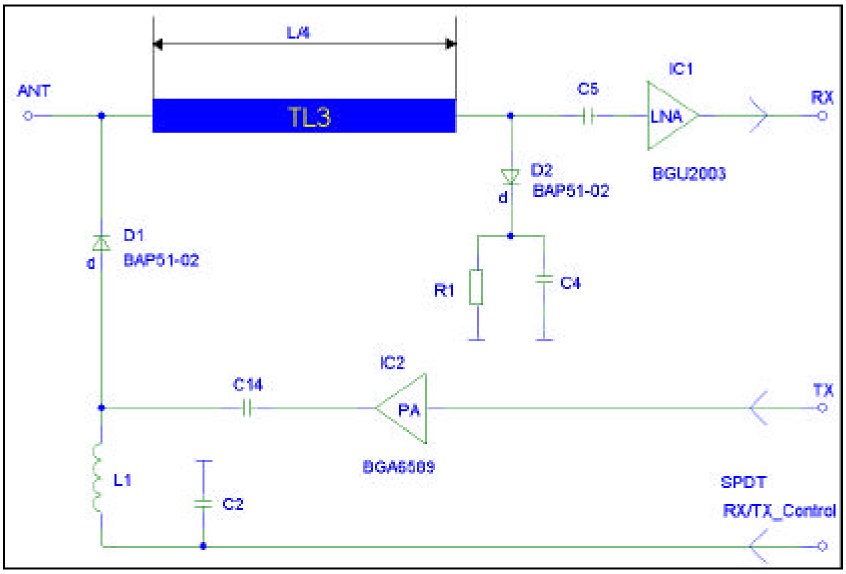


Figure 30: Principle working of the SPDT for multiplexing PA and LNA

A dc voltage on RX/TX Control terminal passes L1 and forward biases the PIN diodes D1 and D2. The dc current is adjusted by R1. Because of the principle function of a PIN diode, forward biased D1 and D2, have a very low resistance R_{ON}. This can be assumed as a RF short. Due to this, the input of the LNA input is shunted via D2 and the capacitor C4 to GND. C5 prevents any change of DC potential at the LNA input. For the principle function, D2 forward biased can be assumed to be a short for RF signals. The result is a very low amount of ANT-Signals amplified by IC1. From the power ratio RX/ANT is calculated the ***RX-ANT isolation*** for switched on transmitter. C14 prevents any dc level change on the PA output.

The mechanical dimensions of the Microstrip (µStrip) transmission line TL3 are designed to be a 50Ω quarter wavelength transformer. That means its electrical length is $L = \frac{l}{4}$. With λ=wave-length inside the used µStrip substrate within the pass band (center frequency). As explained in the RF-Design-Basics chapter, the L/4 line does transform an impedance: $Z_{OUT} = \frac{Z_L^{\,2}}{Z_{IN}}$ A "short" on one side causes the L/4 -transformer a transformation into an "open" appearing on the opposite µStrip side. The mathematical issue is shown side by. Due to this function, the LNA input is shunted to GND. At the opposite side of TL3, the RX-rail is high resistive and can't absorb RF power. That means the RX-rail is switched out of the circuit and only a very low amount of PA power can leak into the LNA. Due to the very low resistant D1, the output power of the PA travels with very low losses to the ANT-terminal. The power ratio of ANT/PA-out is the switch ***TX-insertion loss***.

Microstrip l/4 transformer analysis:

Transmission-Line ***(TL)***: $\underline{Z_1} = Z_L \dfrac{\dfrac{Z_2}{Z_L} + j \tan b \cdot L}{1 + j \dfrac{Z_2}{Z_L} \tan b \cdot L}$

$$\underline{Z_1} = Z_L \frac{\dfrac{Z_2}{Z_L} + j \tan\left(2p \dfrac{L}{l}\right)}{1 + j \dfrac{Z_2}{Z_L} \tan\left(2p \dfrac{L}{l}\right)} = Z_L \frac{x + j \tan\left(2p \dfrac{L}{l}\right)}{1 + jx \tan\left(2p \dfrac{L}{l}\right)}$$

with $L = \dfrac{l}{4}$ causes $\underline{Z_1} = Z_L \dfrac{x + j \tan\left(\dfrac{p}{2}\right)}{1 + jx \tan\left(\dfrac{p}{2}\right)}$

$\tan\left(\dfrac{p}{2}\right) = \infty$ => non defined ratio $\dfrac{\infty}{\infty}$ by lim analysis

With L=length of the Transmission-Line.

Continued on the next page…

The remaining TX signals appearing at the RX output are defined by the power ratio RX/TX and called ***RX/TX coupling***. Removal of the RX/TX Control dc voltage put the PIN-diodes in the off state. In this sate they are highly resistive with a very low junction capacitance. This is another very important characteristic of PIN diodes. In this bias mode the output power of IC2 are blocked by D1 and can't reach the ANT-terminal ***(TX-PA isolation or TX leakage)***. Because D2 is very high resistive, the µStrip does only see the LNA's input impedance of 50Ω. As illustrate by the L/4 mathematical evaluation, the µStrip output impedance will be the same as offered on the opposite side about 50Ω. Due to it, the ANT-signals are low loosely transferred to the LNA and appear low noise amplified at the RX output terminal. The diodes D1 and D2 do form a switch with one common PIN and two independent pins. This construction is called a single pole double trough switch ***(SPDT)***.

$$\Rightarrow \underline{Z_1} = Z_L \frac{x + jy}{1 + jxy} = Z_L \frac{\frac{x}{y} + j}{\frac{1}{y} + jx}$$

$$\lim_{y \to \infty} \left(Z_1 \right) = Z_L \frac{0 + j}{0 + jx} = Z_L \cdot \frac{j}{jx}$$

$$\underline{Z_1} = Z_L \cdot \frac{Z_L}{Z_2} \quad \Rightarrow \quad \boxed{\underline{Z_1} = \frac{Z_L^{\,2}}{Z_2}}$$

Special cases:

open	==>	short;	C	==>	L
short	==>	open;	L	==>	C

1.2.5.2. Circuit Details

- ***PLEASE NOTE: - DC SUPPLY SETUP -***

For protecting the Reference Board against over voltage and wrong polarity during bench experiments, the main board connectors do have an input shunt Z-Diode {D7, D8, D9}. In a bias fault condition the diodes shunt the dc terminals to GND. Due to it, please adjust the current limiter of your dc power supply and check out for proper polarity and right amount of dc voltage. Several LEDs on the board monitors the main board functions for visual feedback the actual modes.

- ***SPDT:***

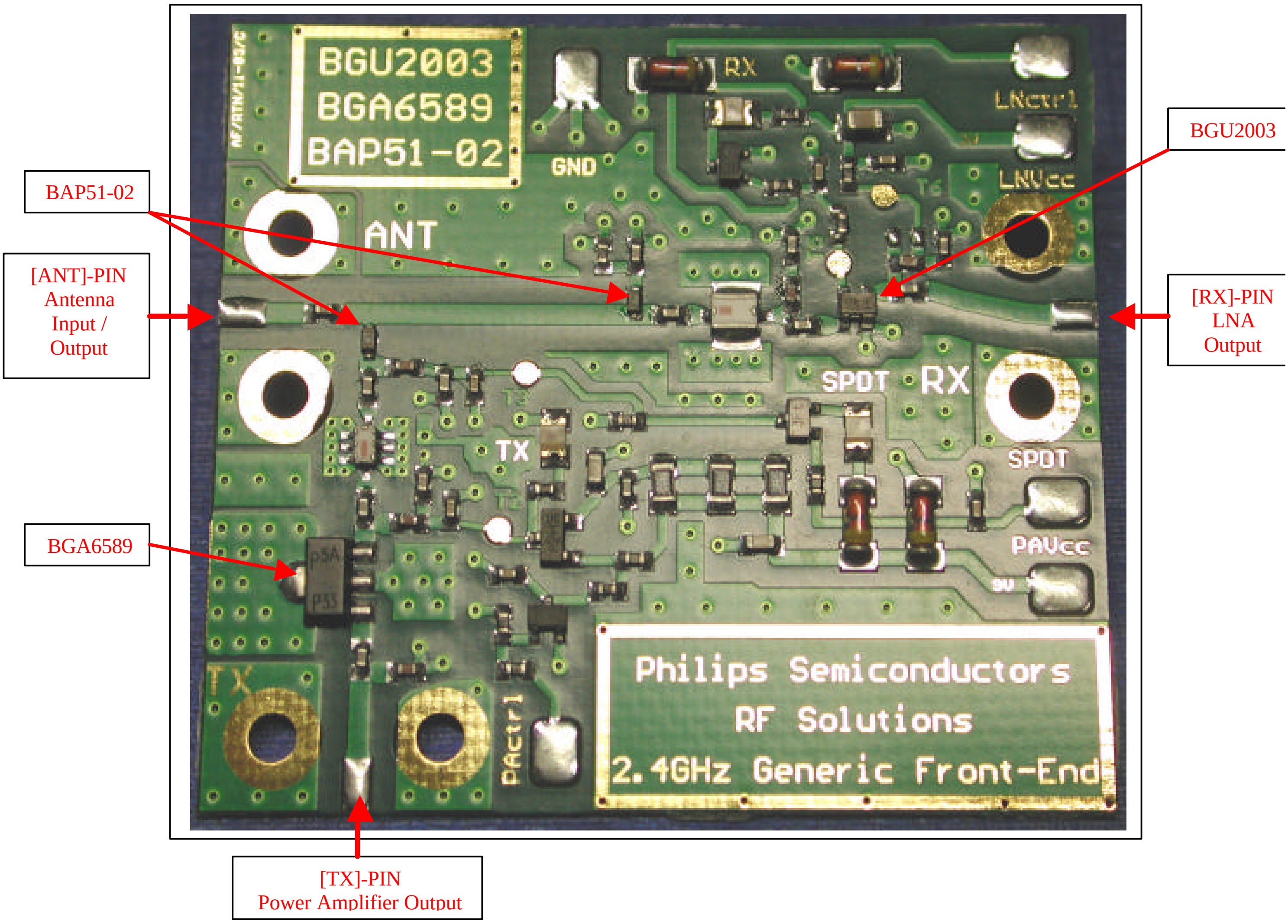
The SPDT switch is build by the circuit {D1, D2, R1, C4, C3, L1, C2, C1}. The circuit Q3, D6, R7, C18, controls the mode of the switch. The PIN diode forward current is set-up by R1. C4 do short the cathode of D2 to GND. C3 couple the Antenna to the switch by removal of dc components. L1 is high resistive for the RF but do pass the dc current into the PIN diodes. C2, C1 do short remaining rests of RF. At Checkpoint T3, the dc voltage across the SPDT switch can be measured. The combination of D6, D5 and B-E junction of Q3 forms a dc level shifter for proper switching of Q3 by a 3V logic signal. A lighting D5 caused by SPDT=LOW do illustrate a SPDT switch mode of the antenna terminal connected to the PA output. C18 removals coupled in of line noise caused by long wires connected to the board. C5, C10 and C14 prevent a dc rail into the MMICs. The principle SPDT function based on the quarter wavelength µStrip line TL3 is explained in the former chapter. Voltages quite below 3V do put the PIN diodes into analog attenuator mode.

- ***LNA:***

The LNA's (IC1) supply bias is comparable to a pull up circuit for an open collector. The LNA supply voltage is connected to terminal LNVcc. C20 and C15 removals switching peaks, coupled-in noise and line growl. D9 do clamp the voltage to abs. max. =3.6V. Input voltage of > 3.6V will source down the current limiter of the used lab power supply. It's for protecting against over voltage and wrong polarity applied to the LNA circuit. R4 do set up the bias operation point of the LNA output circuit. C6 defines a clear short to GND for the L2. L2-C7 combination forms an output L-matching circuit for the LNA. Additionally L2 do dc bias MMIC at PIN4. The optional R3 can be used for making more broadband the output circuit (Q decreased) or for damping of oscillation. The bias point and gain adjust is done by a current into the control PIN3. The control current is adjusted and limited by R2. C16 acts for wire noise reduction. D8 protects again over voltage (>3.6V) and wrong polarity. With LNctrl=HIGH, the LNA is switched on with max. Gain. This is illustrated by lighting D3. LNctrl voltages between 0V and 3.0V can be used for standby, max. Gain and variable gain applications like AGC. The voltage potential difference between LNctrl and test point T5 (across R2) can be used for calculating the actual control current into PIN3. Depending on the amount of R12, the LED D3 do illustrate the actual LNA-Gain. The LNA input impedance and the optimum noise impedance are closed to 50Ω. C5 do removal dc components. The input return loss is optimized by the combination L4-C5 appearing as a resonance match at ANT connector X1.

- ***PA:***

The power amplifier MMIC (IC2) does it self need a supply of ca. 4.7V/83mA sinking into the output PIN3. For temperature stabilization of the output voltage-current temperature relationship, there is the need of series resistors {R8, R13, and R14}. L3 do inject the dc supply current into the MMIC. Additionally L3 blocks the RF. RF leakage behind it is shunt to GND by C11. C12 do back up for medium frequencies and ripples cause by e.g. large output envelope change. At test point T2 can be monitored the PA output dc voltage. By the use of {Q1, R10, C19} the PA can be switched off. Circuit Q4, R16, R17 makes the PActrl connector compatible to standard logic ICs. Depending on the logic output swing, a pull up resistor is need. With PActrl = Logic HIGH, D4 does light indicating switched on power amplifier. L5 does optimize input return loss. C10 prevents the MMICs internal input dc bias shift by circuits connected to X3. D7 do protect the PA against over voltage and wrong polarity applied to the PAVcc connector X5.



1.3. "2.4GHz Generic Front-End Reference Board" Data Sheet

Philips Semiconductors European Support Group

Board specification
2004 January

2.4GHz Generic Front-End Reference Board

BGA6589
BGU2003
BAP51-02

FEATURES

- 2.4GHz ISM band operation
- 50Ω female SMA connectors
- LNA, PA and SPDT on one board
- Supply control function
- LED's indicates the operation mode

APPLICATIONS

- Bluetooth
- W-LAN
- ISM
- Home video and TV link
- Remote control
- Consumer, Industrial, Automotive

DESCRIPTION

The Reference Board is intended to be used as a generic Front-End circuit in front of a high integrated half duplex IC chip set. It uses a LNA: SiGe MMIC amplifier (BGU2003) for improving the receiver's sensitivity and a PA: MMIC wideband medium power amplifier (BGA6589) for increasing the transmitter distance. A digital controlled antenna switch (SPDT): General purpose PIN-Diodes (BAP51-02) for multiplexing the LNA-input or the PA-output to the common antenna terminal (e.g. terminated by a 50Ω ceramic antenna).

PINNING

PIN / PORT	DESCRIPTION
ANT	Bi-directional common 50Ω Antenna I/O
GND	Ground
LNctrl	LNA control input
LNVcc	LNA dc supply
RX	LNA 50Ω output
SPDT	SPDT control RX/TX
PAVcc	PA dc supply
PActrl	PA control input
TX	PA 50Ω input

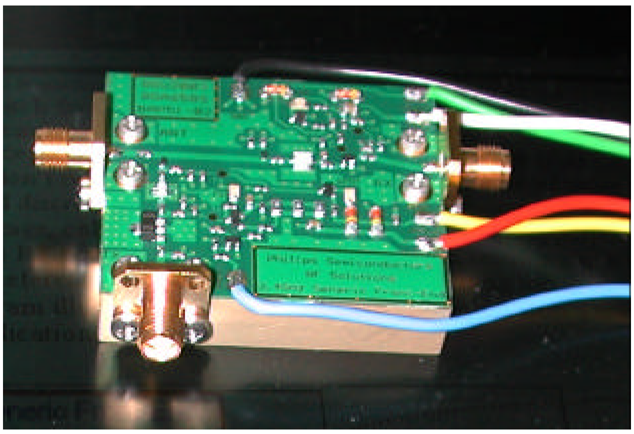


Figure1: Reference Board Rev. C Top View

QUICK REFERENCE DATA

PAVcc=9V; LNVcc=3V

SYMBOL	PARAMETER	CONDITIONS	MIN.	TYP.	MAX.	UNIT
BW	bandwidth	Limited by the used filters		2400 to 2500		MHz
PAVcc	DC supply voltage	PA		9		V
LNVcc	DC supply voltage	LNA		3		V
$I_{(PAVcc)}$	supply current power amplifier (PA)	all ports 50Ω terminated; PActrl=3V; SPDT=5V	73,2	83,4	86,8	mA
$I_{(LNVcc)}$	supply current low noise amplifier (LNA)	LNctrl=3V; all ports 50Ω terminated; LNVcc=3V	13,9	16,3	17,7	mA
$I_{(SPDT\text{-}switch)}$	Antenna PIN diode switch (SPDT) bias current	all ports 50Ω terminated SPDT=0V	≈3,1	≈3,2	≈3,7	mA
$I_{(stby)}$	standby supply current	$I_{(PAVcc)} + I_{(LNVcc)}$ SPDT=3V; PActrl=LNctrl=0V	0,8	1,2	1,6	mA
S_{21}	forward power gain	LNA receive (RX); 2450MHz	10,7	12	12,8	dB
		PA transmit (TX) ; 2450MHz	14,2	14,5	14,8	dB
NF	noise figure LNA	PActrl=0V; LNctrl=3V; SPDT=3V; 2400MHz	3,2	3,3	3,5	dB
		2450MHz	3,2	3,3	3,3	dB
		2500MHz	3,3	3,3	3,4	dB
$P_{L\ 1dB}$	output load power at 1dB gain compression	LNA output; 2450MHz; SPDT=3V	+10,5	+11,1	+11,7	dBm
		PA output; 2450MHz; SPDT=0V	+18,3	+18,6	+18,9	dBm
IP3	output 3rd order intercept point 2450MHz+2451MHz	LNA output; LNctrl=SPDT=3V; PActrl=0V	+21,2	+23,1	+24,2	dBm
		PA output; LNctrl=SPDT=0V; PActrl=3V	+31,2	+31,6	+32	dBm
LNctrl	LNA = standby	LNctrl=L		0		V
	LNA = RX operation	LNctrl=H; LNctrl<LNVcc		3		V
PActrl	PA = standby	PActrl=L		0		V
	PA = TX operation	PActrl=H		3 to 5		V
SPDT	ANT connected to TX rail	SPDT=L		0		V
	ANT connected to RX rail	SPDT=H		3 to 5		V

Note:

1. Typically (TYP) data are the average measured over 10 prototype hand made boards Rev. B.
2. MIN and MAX are distribution extreme measured over 10 prototype boards Rev. B
3. PL1 tested with SME03 and hp8594E (int. 40dB attenuator fixed) on 10 prototype boards Rev. B

LIMITING VALUES

SYMBOL	PARAMETER	CONDITIONS	MIN.	MAX.	UNIT
PAVcc	DC supply voltage	see Note 1	0	<10	V
LNVcc	DC supply voltage	see Note 1	0	<3,6	V
SPDT	SPDT switch control		0	PAVcc	V
LNctrl	LNA power control	LNctrl<LNVcc; see Note 1	0	<3,6	V
PActrl	PA power control		0	tbf	V

Note:

1. The boards connectors LNVcc, LNctrl, PAVcc are protected by a Z-Diode to GND. Negative voltages or voltage at the limit do cause the diode shunting a large current to GND. This is for protecting the board against wrong polarity and over voltage during bench experiments.

ACTIVE DEVICES THERMAL CHARACTERISTICS

SYMBOL	PARAMETER	CONDITIONS		VALUE	UNIT
$R_{th\ j\text{-}s}$	thermal resistance from junction to solder point	BGA6589, $T_S \leq 70$ °C; note 1		100	K/W
		BGU2003		85	
		BAP51-02		350	
$R_{th\ j\text{-}a}$	thermal resistance from junction to ambient	BC847BW; note 2		625	
		BC857BW; note 3		625	
		PBSS5140T	in free air; note 4	417	
			in free air; note 5	278	

Note:

1. T_S is the temperature at the soldering point of pin 4.
2. Transistor mounted on a FR4 printed-circuit board.
3. Refer to SOT323 standard mounting conditions.
4. Device mounted on a printed-circuit board, single sided copper, tinplated and standard footprint.
5. Device mounted on a printed-circuit board, single sided copper, tinplated and mounting pad for collector 1cm^2.

DETAILED PINNING DESCRIPTION

PIN Name	SYMBOL	NAME AND FUNCTION
ANT	X1	Antenna connector; input for receive (RX); output for transmit (TX); 50Ω; RF bidirectional
RX	X2	RX-Out connector; 50Ω; RF output
TX	X3	TX-IN connector; 50Ω; RF input
LNctrl	X4	Digital input. Supply control for LNA amplifier: H➔LNA=ON; L➔LNA=Standby
PAVcc	X5	+9Vdc; supply voltage for the power amplifier (PA) and for SPDT Antenna switch
GND	X6	0Vdc; common for all functions
SPDT	X7	Digital input. Control signal for the antenna switch: L ➔ X1=PA-TX-Output; X3=PA-TX-Input ➔ Transmit mode H ➔ X1=LNA-RX-Input; X2=LNA-RX-Output ➔ Receive mode
PActrl	X8	Digital input. Supply control for transmit (TX) power amplifier (PA): L ➔ PA=OFF; H ➔ PA=ON
LNVcc	X9	+3Vdc; supply voltage for the low noise amplifier (LNA)

FUNCTIONAL TABLE

Digital logic description

INPUTS			RF-CONNECTORS			ON BOARD LED STATUS CONTROL			FUNCTION
LNctrl	PActrl	SPDT	RX (X2)	TX (X3)	ANT (X1)	D3 (RX)	D4 (TX)	D5 (SPDT)	
A	B	L	Fc	IN	OUT	A	B	H	Antenna connected to TX rail
A	B	H	OUT	Fc	IN	A	B	L	Antenna connected RX rail
H	B	C	IN	Fc	Fc	H	B	$\overline{C}$	LNA amplifier switched on
L	B	C	X	Fc	Fc	L	B	$\overline{C}$	LNA amplifier switched off
A	L	C	Fc	X	Fc	A	H	$\overline{C}$	PA amplifier switched off
A	H	C	Fc	IN	OUT	A	L	$\overline{C}$	PA amplifier switched on

Notes:

1. A, B, C = Variable substituting the logic level. It can be L or H steady state.
2. Fc = Function not changed
3. L = Low voltage level steady state; LED=off
4. H = High voltage level steady state; LED=on
5. IN = Connector works as an input
6. OUT = Connector works as an output
7. D3-D5 = On board LED status. LEDs do have the labels RX, TX, ANT
8. TX rail = Transmitting circuit of the reference board
9. RX rail = Receiving circuit of the reference board

Mathematical description of the digital functions:

$$\mathrm{RXmode} = \mathrm{LNctrl} \wedge \mathrm{SPDT}$$

$$\mathrm{TXmode} = \mathrm{PActrl} \wedge \overline{\mathrm{SPDT}}$$

$$\mathrm{D3} = \mathrm{LNctrl}$$

$$\mathrm{D4} = \mathrm{PActrl}$$

$$\mathrm{D5} = \overline{\mathrm{SPDT}}$$

DC LEVELS OF THE LOGIC SIGNALS

SYMBOL	PARAMETER	CONDITIONS	MIN.	TYP.	MAX.	UNIT
LNctrl	LNA = off = standby	LNctrl=L		0		V
	LNA = on = RX operation	LNctrl=H		3		V
PActrl	PA = off = standby	PActrl=L		0		V
	PA = on = TX operation	PActrl=H		3		V
SPDT	ANT connected to TX rail	SPDT=L		0		V
	ANT connected to RX rail	SPDT=H		3		V

CHARACTERISTICS DATA DEFINITION

The MIN. and MAX. data are the data spread measured on 10 investigated handmade prototype boards. The TYP. data is arithmetic average of the measurement done on this boards. The LSL and USL are the final test limits. Not all parameters measured on the prototype boards were tested on the machine-manufactured batch of 120 boards. If a parameter (SYMBOL) is tested during final test, than LTL and/or UTL are specified. In this case, the MIN., MAX. and TYP. fields do list the test results found on the machine manufactured 120 board batch .

Note:

1. LTL and UTL are the final test limits.
2. LTL = Lower Test Limit for Final-Test
3. UTL = Upper Test Limit for Final-Test
4. MIN. = Minimum data distribution measured found on 10 tested prototype boards
5. MAX. = Maximum data distribution measured found on 10 tested prototype boards
6. TYP.=Calculated average of the data distribution measured on 10 prototype boards
7. Good boards (BIN1) must be within the final test limits (LTL ≥ pass ≤ UTL)
8. If data fields LTL or UTL are empty (---), this parameter (symbol) will not be Final-Tested or do not have this limit.
9. The data MIN, MAX, TYP were measured at 2401MHz, 2449.75MHz and 2498.5MHz. This is, because of done test over broadband frequency range, causing limitation frequency resolution of the Network Analyzer (ZVRE). Final-Test should be done at the integer frequencies 2400, 2450 and 2500MHz.
10. The Reference Board's Data Sheet does not expand, limit or influence the data sheets of the used parts.

STATIC CHARACTERISTICS

PAVcc=9V; LNVcc=3V; Tj=room temperature; all ports 50Ω terminated;
unless otherwise specified

SYMBOL	PARAMETER	CONDITIONS	LTL	MIN.	TYP.	MAX.	UTL	UNIT
$I_{(LNVcc)}$	supply current LNA	LNctrl=0V; LNA=off		0,8	1,1	1,5	1,7	mA
		LNctrl=3V; LNA=on	11,6	13,9	16,3	17,7	19,4	mA
$I_{(PAVcc)}$	supply current PA=off	SPDT=5V; PActrl=0V		0	0,8	1,1		µA
		SPDT=3V; PActrl=0V		43	56,1	66	66	µA
		SPDT=0,5V; PActrl=0V		2	2,9	3,3		mA
		SPDT=0V; PActrl=0V	2,3	3	3,2	3,7	3,8	mA
	supply current PA=on	SPDT=5V; PActrl=3V	64	73,2	83,4	86,8	108	mA
LNctrl	LNA = standby	LNctrl=L			0			V
	LNA = RX operation	LNctrl=H; LNctrl<LNVcc			3	LNVcc		V
PActrl	PA = standby	PActrl=L			0			V
	PA = TX operation	PActrl=H			3 to 5			V
SPDT	ANT connected to TX rail	SPDT=L; ANT=$PA_{(OUT)}$			0			V
	ANT connected to RX rail	SPDT=H; ANT=$LNA_{(IN)}$			3 to 5			V

Note:
Their were investigated several standard logic families and microcontrollers in different technologies operating at different supply voltages. Typically PActrl and SPDT do identify a logic state level High at +3V. Increasing up to 5V (TTL standard logic) is possible and can slightly improve some parameters of the power amplifier and of the antenna switch. Under load, real logic ICs often doesn't offer rail-rail output swing. If the output logic High level gets critically, a pull-up resistor may help. Typically logic low sate of isn't critically. Alternatively the resistors in the digital part of the Reference Board must be changed or e.g. an use off an additionally external open collector transistor do help. Philips Semiconductors open collector comparator amplifiers like NE522 or rail-to-rail operational amplifier family NE5230 or NE5234 may be interesting for certain applications too.

CHARACTERISTICS: Return Loss of the Transmitter

PAVcc=9V; LNVcc=3V; RX=50Ω terminated; LNctrl=3V=on; Tj=room temperature; unless otherwise specified

SYMBOL	PARAMETER	CONDITIONS		LTL	MIN.	TYP.	MAX.	UTL	UNIT
RL $_{IN\ TX}$	return loss input TX; PA=off	SPDT=0V PActrl=0V SPDT=TX	2400MHz		4	4,4	4,7		dB
			2450MHz		4	4,4	4,7		
			2500MHz		4	4,4	4,7		
	return loss input TX; PA=on	SPDT=0V PActrl=3V SPDT=TX	2400MHz		13	14,3	16,1		dB
			2450MHz		13,1	14,4	16,3		
			2500MHz		13,1	14,4	16,5		
		SPDT=3V PActrl=3V SPDT=RX	2400MHz		15,9	17,3	19,9		dB
			2450MHz		17,9	19	20,3		
			2500MHz		19,3	20,5	21,5		
RL $_{OUT\ ANT}$	return loss output ANT; PA=on	SPDT=0V PActrl=3V SPDT=TX	2400MHz		10,9	13,2	15,1		dB
			2450MHz	9,5	12	13,7	16,4		
			2500MHz		9,8	11,9	13,6		
	return loss output ANT; PA=off	SPDT=0V PActrl=0V SPDT=TX	2400MHz		8,8	9,9	10,8		dB
			2450MHz		8,2	9,4	10,4		
			2500MHz		7,7	9	10,1		
		SPDT=1V PActrl=0V SPDT=RX	2400MHz		4,1	9,4	10,7		dB
			2450MHz		4,1	9	10,3		
			2500MHz		4,1	8,6	10		

Note:

1. NWA=Network Analyzer, source power -30dBm at both test ports (20dB-step attenuator; -10dBm-source)

PAVcc=9V; LNVcc=3V; RX=50Ω terminated; LNctrl=0V=off; Tj=room temperature; unless otherwise specified

SYMBOL	PARAMETER	CONDITIONS		LTL	MIN.	TYP.	MAX.	UTL	UNIT
RL $_{IN\ TX}$	return loss input TX; PA=off	SPDT=0V PActrl=0V SPDT=TX	2400MHz		3	4,5	4,7		dB
			2450MHz		4	4,4	4,7		
			2500MHz		4	4,4	4,7		
	return loss input TX; PA=on	SPDT=0V PActrl=3V SPDT=TX	2400MHz	11	12,3	14,2	17,8		dB
			2450MHz	11,5	12,4	14,3	16,9		
			2500MHz	11,5	12,5	14,4	16		
		SPDT=3V PActrl=3V SPDT=RX	2400MHz		15,8	17,2	19,9		dB
			2450MHz		17,8	18,9	20,3		
			2500MHz		19,4	20,3	21,3		
RL $_{OUT\ ANT}$	return loss output ANT; PA=on	SPDT=0V PActrl=3V SPDT=TX	2400MHz	10	11,9	13,3	15,7		dB
			2450MHz	9,5	11,3	12,6	15		
			2500MHz	9,2	10,8	11,9	13,2		
	return loss output ANT; PA=off	SPDT=0V PActrl=0V SPDT=TX	2400MHz		4	8,3	10,5		dB
			2450MHz		4	8,4	10,2		
			2500MHz		4	8,1	9,9		
		SPDT=1V PActrl=0V SPDT=TX	2400MHz		8,8	9,3	10,6		dB
			2450MHz		4,6	8,9	10,2		
			2500MHz		4,7	8,6	9,9		

CHARACTERISTICS: Return Loss of the Receiver

PAVcc=9V; LNVcc=3V; TX=50Ω terminated; LNctrl=3V=on; Tj=room temperature;
unless otherwise specified

SYMBOL	PARAMETER	CONDITIONS		LTL	MIN.	TYP.	MAX.	UTL	UNIT
$RL_{IN\ ANT}$	return loss input ANT; PA=off	SPDT=3V LNctrl=3V PActrl=0V	2400MHz	8	8	11,5	20,8		dB
			2450MHz	8	12,2	16,9	26,4		
			2500MHz	8	8,6	15	28		
	return loss input ANT; PA=on	SPDT=0V LNctrl=0V PActrl=3V	2400MHz		9,7	12,7	15		dB
			2450MHz		9,3	11,9	14,1		
			2500MHz		9	11,4	13,4		
		SPDT=3V LNctrl=3V PActrl=3V	2400MHz		10,2	12,1	14,2		dB
			2450MHz		15,4	20,7	31,8		
			2500MHz		12,3	15,4	21,4		
$RL_{OUT\ RX}$	return loss output RX; PA=off	SPDT=3V LNctrl=3V PActrl=0V	2400MHz	8	8	9,6	16,9		dB
			2450MHz	8	9,8	11,8	18,5		
			2500MHz	8	11,2	14,4	22,2		
	return loss output RX; PA=on	SPDT=0V LNctrl=0V PActrl=3V	2400MHz		3,4	4,5	12		dB
			2450MHz		3,1	4,3	11,2		
			2500MHz		2,8	4	10,6		
		SPDT=3V LNctrl=3V PActrl=3V	2400MHz		10,4	13,3	19,5		dB
			2450MHz		13,6	16,2	20,5		
			2500MHz		12,3	18,2	22		

CHARACTERISTICS: RX &TX gain, coupling

PAVcc=9V; LNVcc=3V; Tj=room temperature; unless otherwise specified
$S21_{(TX)}$: NWA Port1-IN TX; NWA Port2-ANT;RX=50Ω matched
$S21_{(TX/RX)}$: NWA Port1-IN TX; Port2-Out RX; ANT=50Ω
$S12_{(TX)}$: NWA Port1-IN TX; NWA Port2-ANT;RX=50Ω matched

SYMBOL	PARAMETER	CONDITIONS		LTL	MIN.	TYP.	MAX.	UTL	UNIT
$S21_{(TX)}$	forward gain PA PA=on/off	SPDT=0V LNctrl=0V PActrl=0V	2400MHz		-18,8	-19,1	-19,5		dB
			2450MHz		-18,7	-19,1	-19,4		
			2500MHz		-18,8	-19,1	-19,5		
		SPDT=0V LNctrl=0V PActrl=3V	2400MHz	13,2	14,2	14,8	15	16,2	dB
			2450MHz	13	14,1	14,6	14,8	15,9	
			2500MHz	12,8	13,8	14,4	14,6	15,7	
$S12_{(TX)}$	reverse gain PA PA=on	SPDT=0V LNctrl=0V PActrl=0V	2400MHz		-18,6	-21,1	-24,8		dB
			2450MHz		-18,6	-21,1	-24,7		
			2500MHz		-18,7	-21,1	-24,6		
$S21_{(RX)}$	forward gain LNA PA=off	SPDT=3V LNctrl=3V PActrl=0V	2400MHz	9,9	9,9	11,8	12,4	13,4	dB
			2450MHz	10,1	10,3	12,2	12,6	13,8	
			2500MHz	10	10,3	12,1	12,6	13,6	
	forward gain LNA PA=on	SPDT=3V LNctrl=3V PActrl=3V	2400MHz		10,4	11,6	12,8		dB
			2450MHz		10,7	11,9	12,8		
			2500MHz		10,5	11,8	12,5		
$S12_{(RX)}$	reverse gain PA PA=on	SPDT=3V LNctrl=3V PActrl=3V	2400MHz		-20,7	-21,3	-21,9		dB
			2450MHz		-20,1	-20,7	-21,1		
			2500MHz		-19,9	-18,8	-20,9		
$S21_{(TX/RX)}$	coupling TX➔RX PA=LNA=on	SPDT=3V LNctrl=3V PActrl=3V	2400MHz		5,8	7,4	8,7		dB
			2450MHz		4,1	7,6	8,5	9,5	
			2500MHz		5	6,7	7,9		

CHARACTERISTICS: LNA out of band gain

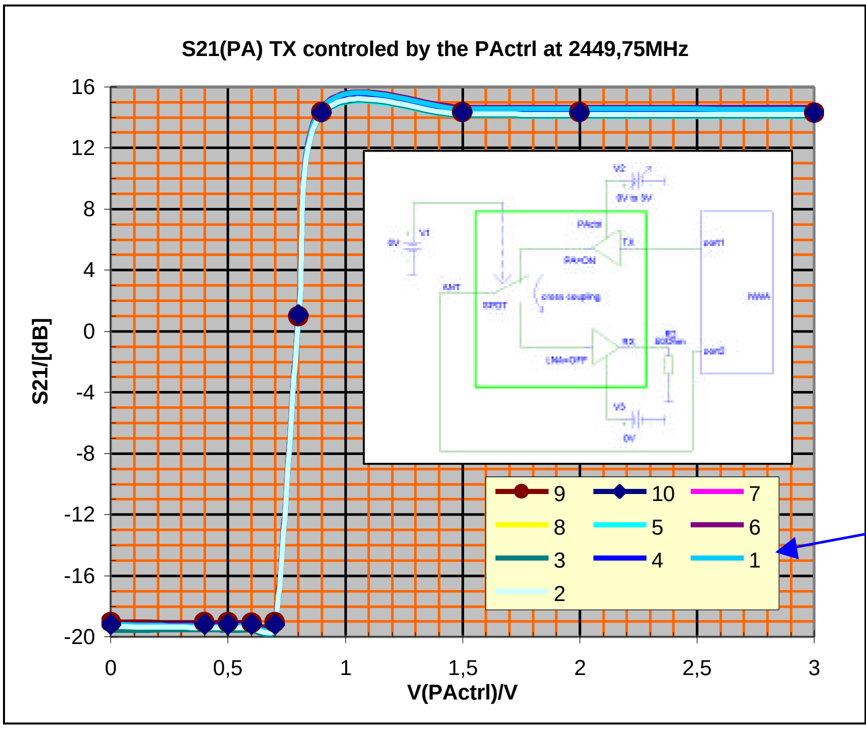
For characterization the sensitivity against received signals out side the 2.4GHz ISM band.

PAVcc=9V; LNVcc=3V; PActrl=0V; LNctrl=SPDT=3V; TX=50Ω matched; Tj=room temperature; unless otherwise specified

SYMBOL	PARAMETER	CONDITIONS	LTL	MIN.	TYP.	MAX.	UTL	UNIT
$S21_{(RX)}$	forward gain LNA	148,71MHz			≈-70			dB
		314,5MHz		-58	-60,7	-65		
		431,5MHz		-50	-52,2	-56		
		899,5MHz		-37,4	-40,2	-44		
		1903,75MHz	-17	-17,7	-25	-31,1		
		2449,75MHz		11,2	12	12,8		
		3600,25MHz	-7,5	-7,5	-8,9	-10,7		
		4000MHz		-16,8	-17,9	-19,9		
		5200MHz		-18	-27,4	-30		
		5800MHz		-20	-24,7	-26,6		

TYPICAL PERFORMANCE CHARACTERISTICS

Performed on 10 hand made prototype boards Rev. B; LNVcc=3V; PAVcc=9V; unless otherwise specified; Tj=room temperature

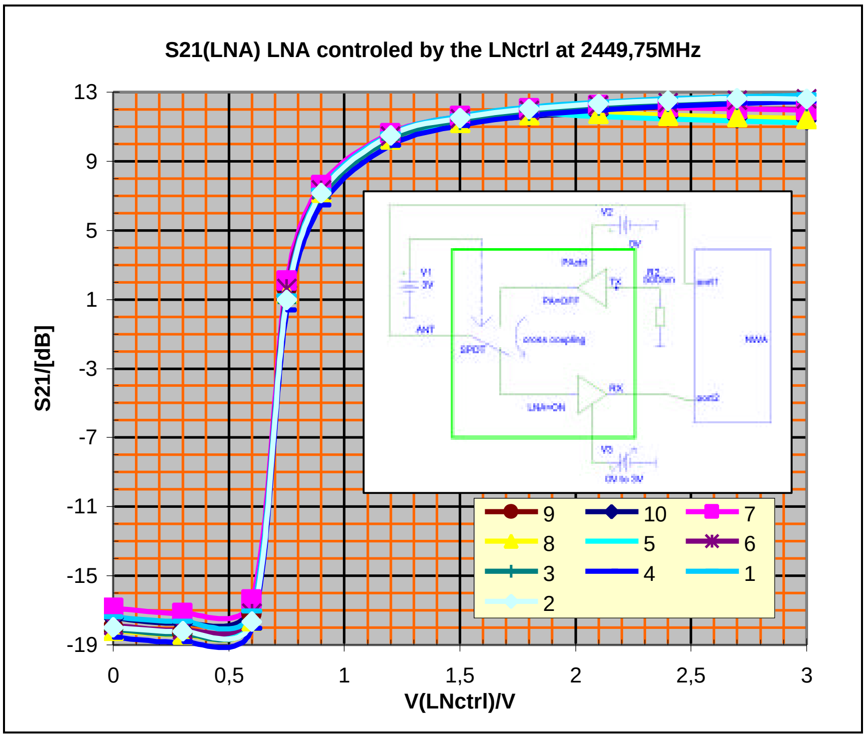


Port2=ANT;Port1=TX; RX=Match

Pactrl=var.; SPDT=0V ; LNctrl=0V

The PActrl pin does control a transistor series connected between PA's Vcc and the supply rail PAVcc. A variation of PActrl does mean a variation of the PA's supply voltage

Prototype Board No #...

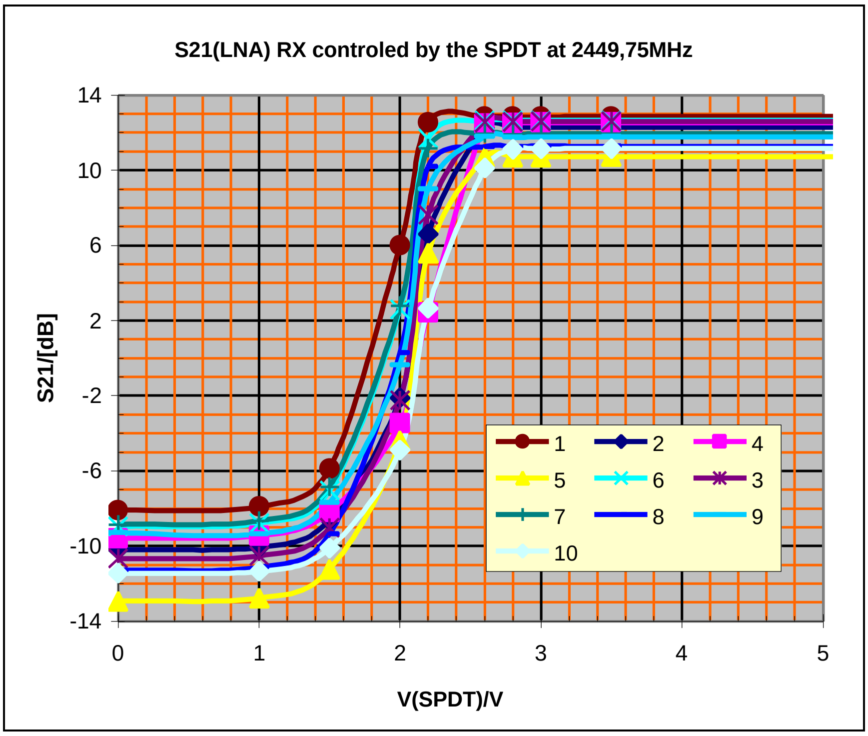


Port2=RX; Port1=ANT; TX=Match

LNctrl=var.; SPDT=3V ; PActrl=0V

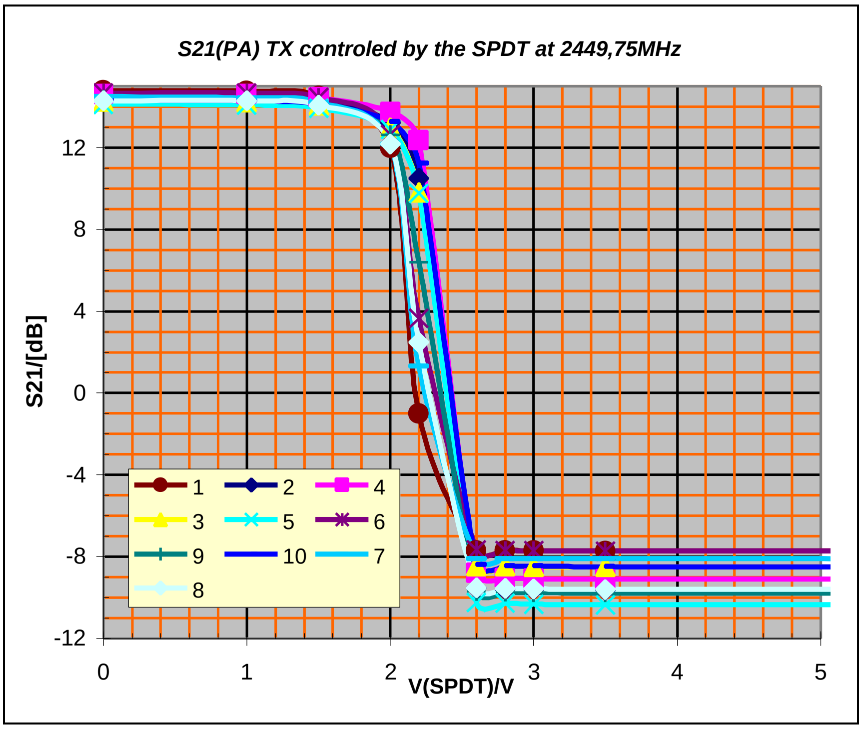
TYPICAL PERFORMANCE CHARACTERISTICS (Continued)

Performed on 10 hand made prototype boards Rev. B; LNVcc=3V; PAVcc=9V; unless otherwise specified; Tj=room temperature



Port2=RX;Port1=ANT; TX=Match

Pactrl=0V; SPDT=VAR; LNctrl=3V

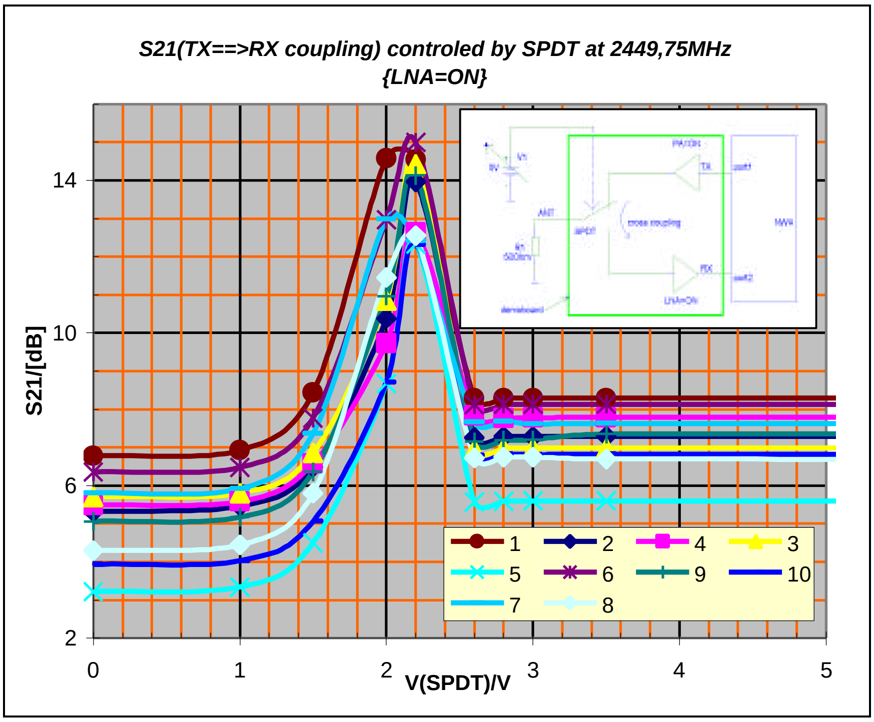


Port2=ANT;Port1=TX; RX=Match

Pactrl=3V; SPDT=VAR ; LNctrl=0V

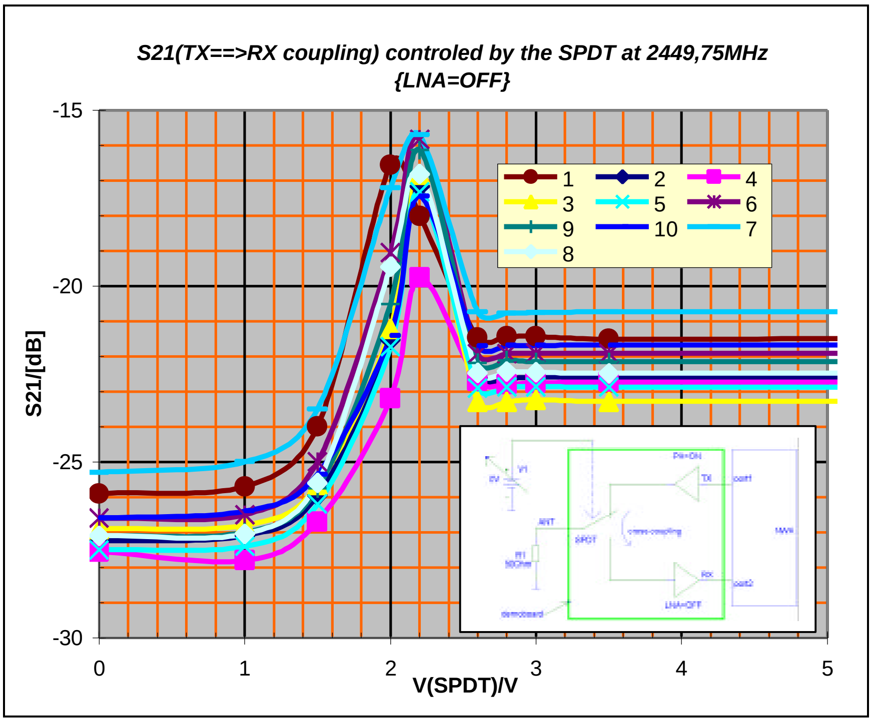
TYPICAL PERFORMANCE CHARACTERISTICS (Continued)

Performed on 10 hand made prototype boards Rev. B; LNVcc=3V; PAVcc=9V; unless otherwise specified; Tj=room temperature



Port2=RX; Port1=TX; ANT=Match

Pactrl=3V; SPDT=VAR ; LNctrl=3V

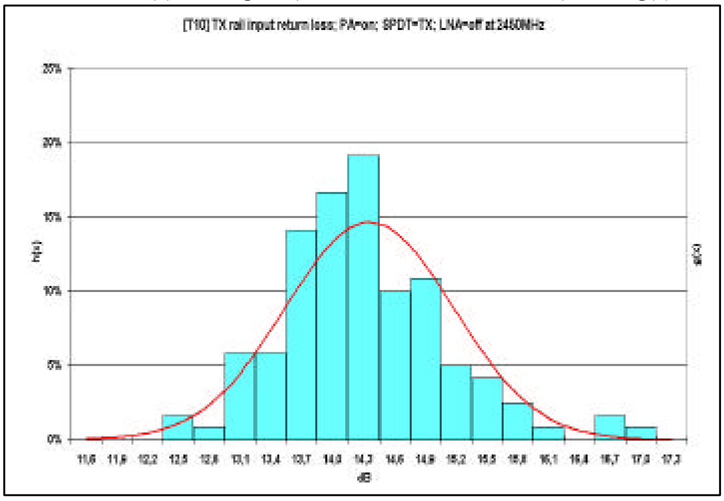


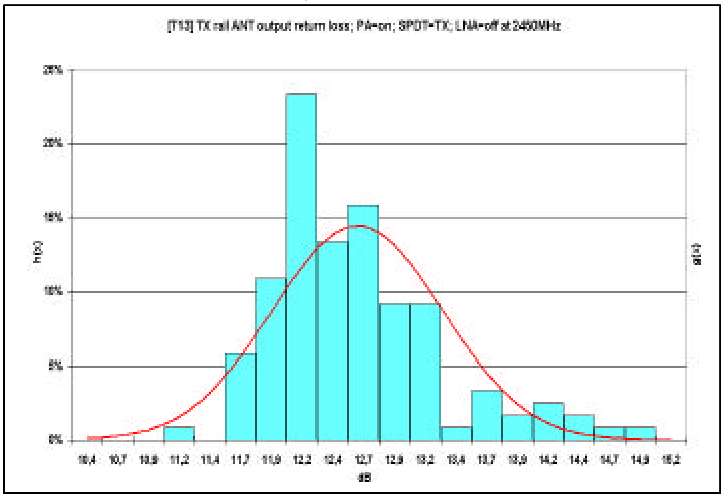
Port2=RX; Port1=TX; ANT=Match

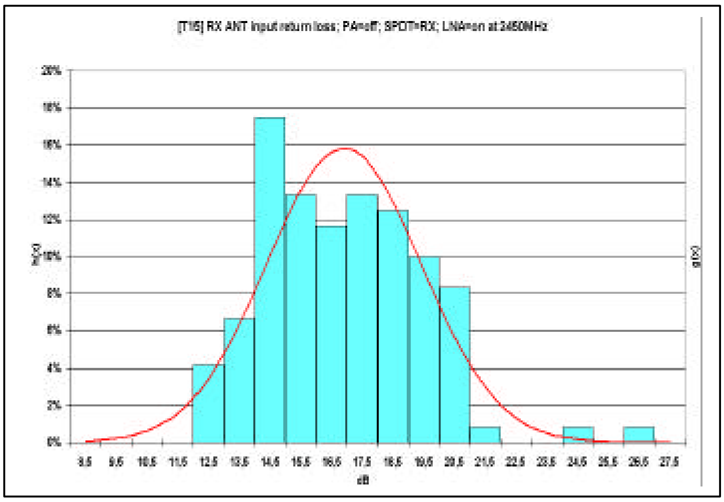
Pactrl=3V; SPDT=VAR ; LNctrl=0V

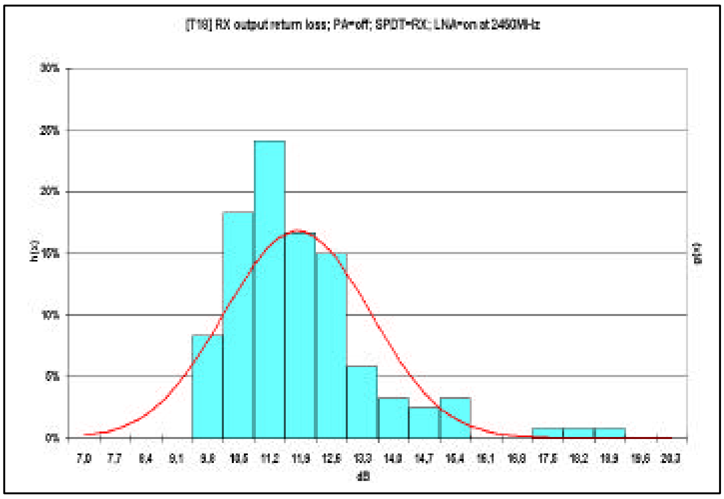
TYPICAL PERFORMANCE CHARACTERISTICS: STATISTICALLY DISTRIBUTION ANALYSIS

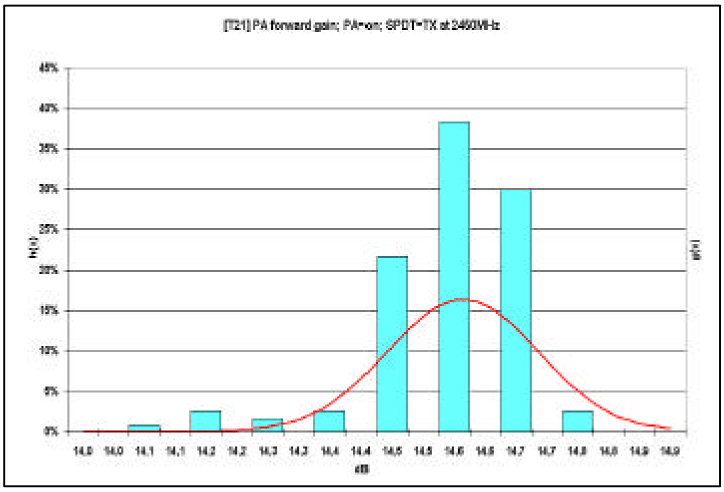
Statistic performed on 120 automatically machine manufactured boards Rev. C; L=0V; H=3V; LNVcc=3V; PAVcc=9V; Tj=room temperature
Blue = h(x) = Histogram (real measured data distribution); Red = g(x) = Normal Distribution (ideal mathematically data evaluation)

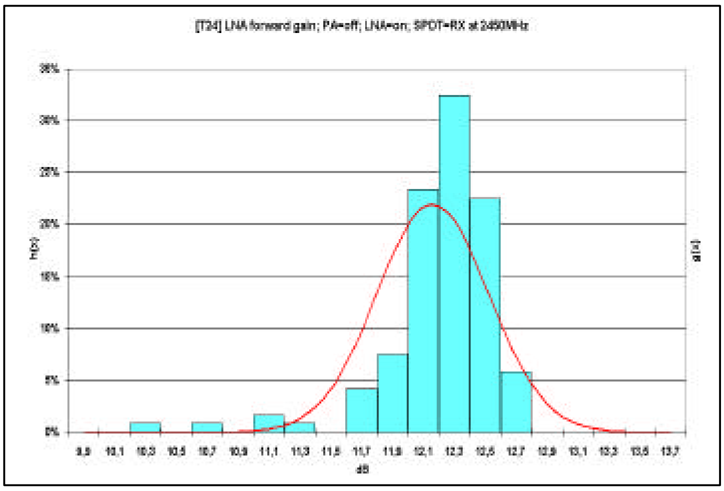






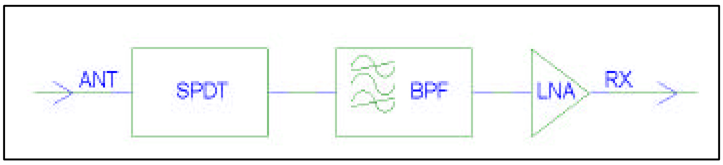






Interpretation of the measured Noise Figure performance

BGU2003's data sheets does list a noise figure of 1.3dB @ 2500MHz. May the reader does ask, why does the board have an effective noise figure of approximately 3.3dB in the receiving rail and whether its useable for his application.



The block diagram illustrates the major noise blocks of the RX rail. It should be take into account, that the noise figure of a passive element is equal to it's insertion loss. From the BGU2003 data sheet first study, can be expect a NF=1.3dB@2500MHz.

Because the demo board's design goal was a good gain and return loss, the realistic NF can be increased. The BGA2003 gain is found out of the S-parameter listing with |S21|=4.325@2500MHz. Philips' AN10173-01, do list for a BAP51-02 based SPDT switch an insertion loss of <0.65dB. For the used RX-band pass filter, the manufacturer does list a max. insertion loss of 1.8dB. These data are taken for doing the following system noise figure analysis on the reference board:

System Noise Figure Factor calculated with **Friis'** noise equitation: $F_g = F_{SPDT} + \frac{F_{BPF} - 1}{G_{SPDT}} + \frac{F_{LNA} - 1}{G_{SPDT} \cdot G_{BPF}}$

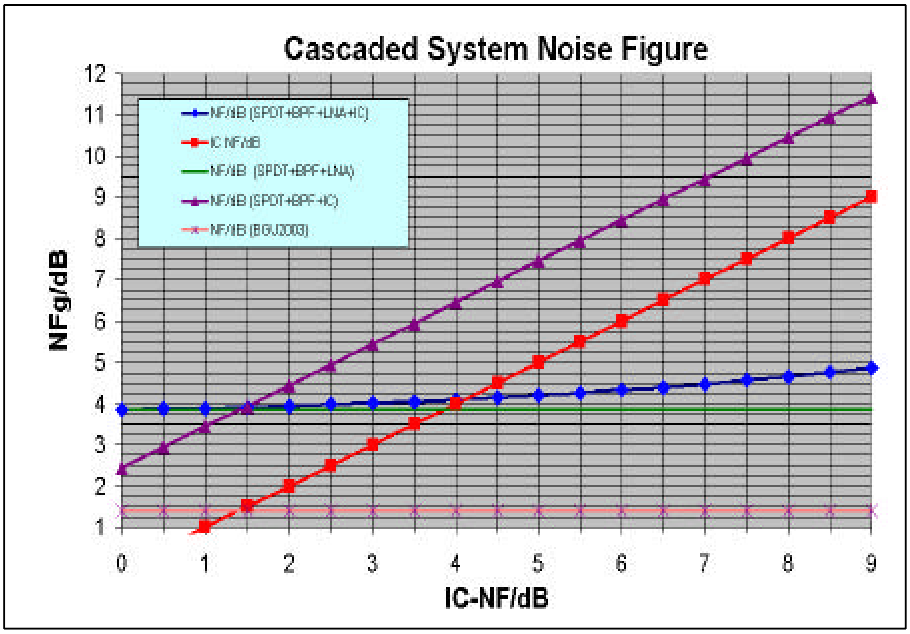
$LNA = 10\log(|S21|^2) = 12{,}72dB$ $\quad G_{SPDT} = 10^{\frac{-0{,}65dB}{10}} = 0{,}861$ $\quad G_{BPF} = 10^{\frac{-1{,}8dB}{10}} = 0{,}661$

$G_{LNA} = 10^{\frac{+12{,}72dB}{10}} = 18{,}71$ $\quad F_{SPDT} = 10^{\frac{0{,}65dB}{10}} = 1{,}161$ $\quad F_{BPF} = 10^{\frac{1{,}8dB}{10}} = 1{,}514$

$F_{LNA} = 10^{\frac{1{,}4dB}{10}} = 1{,}38$ $\quad F_g = 1{,}161 + 0{,}597 + 0{,}668 = 2{,}43$ ➔ $NF = 10\log(F_g) = 3{,}85dB$

The cascaded gain: $L_g = L_{SPDT} + L_{BPF} + L_{LNA} = 10{,}3dB$

The mathematically solving shows a larger board NF than practically measured. Lower insertion loss of the filter and from the SPDT switch combined with a lower NF of the LNA may be the rood cause. Measurements on 10 investigated prototypes showed an average RX gain ≈12dB. If there is an anomaly reading of NF or gain by the noise figure analyzer, a 6dB attenuator between the RX-output and the NF-meter input may help, because a **Yig**-filter (**Y**ttrium-**I**ron-**G**arnet) in the NF-Analyzer input can be very mismatched out of its pass band. Additionally, customer can experiment with the optional resistor R5 or the LNA's output matching circuit depending on the needs of his final application circuit.



The diagram illustrates NF system analysis done on Front-End Reference Board based on Friis' noise law versus the noise figure of an IC chipset. The trade off is approximately 1,5dB. That means IC chipsets (red trace) with NF>1,5dB can be improved by the use of the Ref. Board (blue trace). ICs with a NF<1,5dB will see the advantage of additionally high linear gain including front end selectivity. The dark violet curve illustrates a chipset with SPDT and band pass filter, but without the LNA. The blue trace illustrates the resulting performance of the Reference Board with example IC chipset (NF=9dB). Very clear is illustrate the advantage of BGU2003 from the violet curve comparing to the blue one. The green trace illustrates the theoretically Noise Figure of the Reference Board (≈3,85dB) itself. At (IC-NF)=0dB can be found on the dark violet curve the effective NF of the Reference Board's passive RX components ≈2,5dB.

1.4. Reference

Author:
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RF Discretes Small Signal Application Engineer

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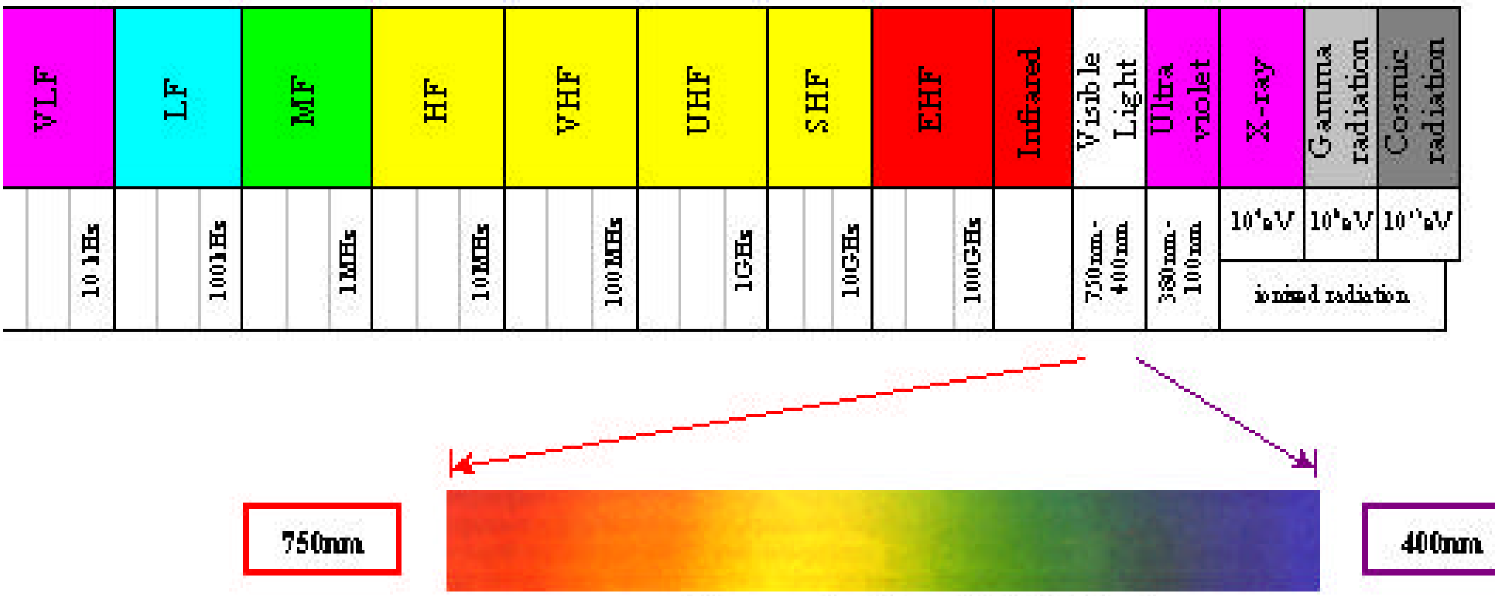
25) UMTS Technik FAQ, www.senderlisteffm.de/umtsfaq.html

Appendix B: RF Application-basics

1.1 Frequency spectrum

Radio spectrum and wavelengths
Each material's composition creates a unique pattern in the radiation emitted.
This can be classified in the "frequency" and "wavelength" of the emitted radiation.
As electro-magnetic (EM) signals travel with the speed of light, they do have the character of propagation waves.



Colour scale of the visible light for human

A survey of the frequency bands and related wavelengths:

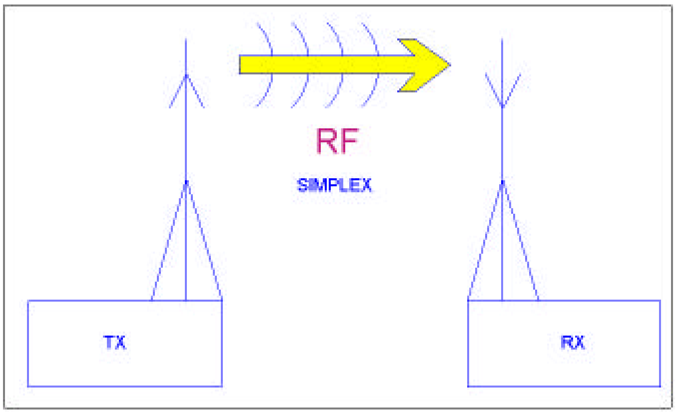
Band	Frequency	Definition (English)	Definition (German)	Wavelength - l acc. DIN40015	CCIR Band
VLF	3kHz to 30kHz	Very Low Frequency	Längswellen (Myriameterwellen)	100km to 10km	4
LF	30kHz to 300kHz	Low Frequency	Langwelle (Kilometerwellen)	10km to 1km	5
MF	300kHz to 1650kHz	Medium Frequency	Mittelwelle (Hektometerwellen)	1km to 100m	6
	1605KHz to 4000KHz	Boundary Wave	Grenzwellen		
HF	3MHz to 30MHz	High Frequency	Kurzwelle (Dekameterwellen)	100m to 10m	7
VHF	30MHz to 300MHz	Very High Frequency	Ultrakurzwellen (Meterwellen)	10m to 1m	8
UHF	300MHz to 3GHz	Ultra High Frequency	Dezimeterwellen	1m to 10cm	9
SHF	3GHz to 30GHz	Super High Frequency	Zentimeterwellen	10cm to 1cm	10
EHF	30GHz to 300GHz	Extremely High Frequency	Millimeterwellen	1cm to 1mm	11
---	300GHz to 3THz	---	Dezimillimeterwellen	1mm-100µm	12

Literature researches according to the Microwave's sub-bands showed a lot of different definitions with very few or none description of the area of validity. Due to it, the following table will try to give an overview but can't act as a reference.

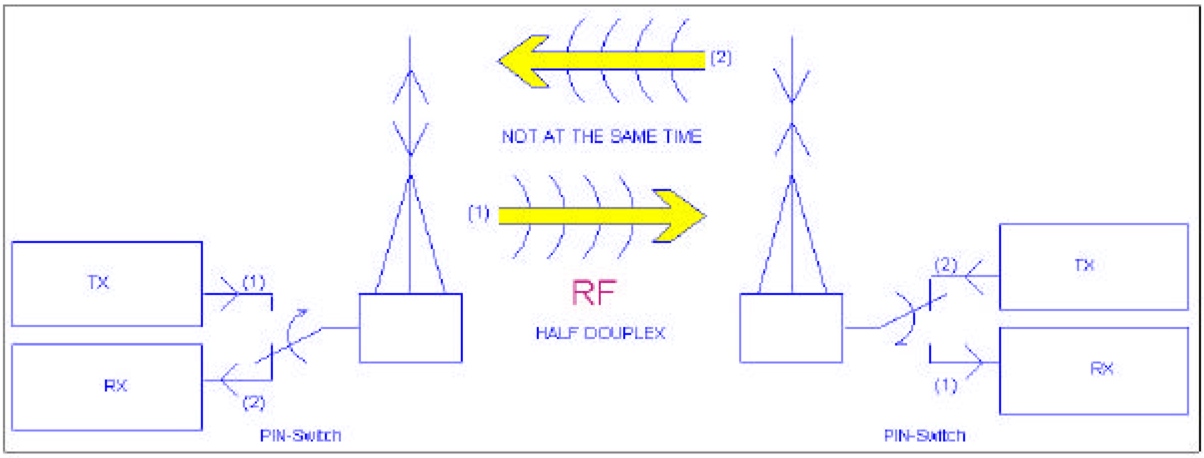
Source	Nührmann	Nührmann	www.wer-weiss-was.de	www.atcnea.de	Siemens Online Lexicon	Siemens Online Lexicon	ARRL Book No. 3126	Wikipedia
Validity	IEEE Radar Standard 521	US Military Band	Satellite Uplink	Primary Radar	Frequency bands in the GHz Area	Microwave bands	---	Dividing of Sat and Radar techniques
Band	GHz	GHz	GHz	GHz	GHz	GHz		GHz
A						0,1-0,225		
C	4-8		3,95-5,8	5-6	4-8	4-8	4-8	3,95-5,8
D		1-3						
E		2-3					60-90	60-90
F		2-4					90-140	
G		4-6					140-220	
H		6-8						
I		8-10						
J		10-20	5,85-8,2					5,85-8,2
K	18-27	20-40	18,0-26,5		18-26,5	10,9-36	18-26.5	18-26,5
Ka	27-40				26,5-40	17-31	26.5-40	26,5-40
Ku	12-18			≈16	12,6-18	15,3-17,2	12.4-18	12,4-18
L	1-3	40-60	1,0-2,6	≈1,3	1-2	0,39-1,55	1-2	1-2,6
M		60-100						
mm	40-100							
P			12,4-18,0			0,225-0,39	110-170	0,22-0,3
R			26,5-40,0					
Q						36-46	33-50	33-50
S	3-4		2,6-3,95	≈3	2-4	1,55-3,9	2-4	2,6-3,95
U			40,0-60,0				40-60	40-60
V						46-56	50-75	50-75
W							75-110	75-110
X	8-12		8,2-12,4	≈10	8-12,5	6,2-10,9	8-12.4	8,2-12,4

1.2 RF transmission system

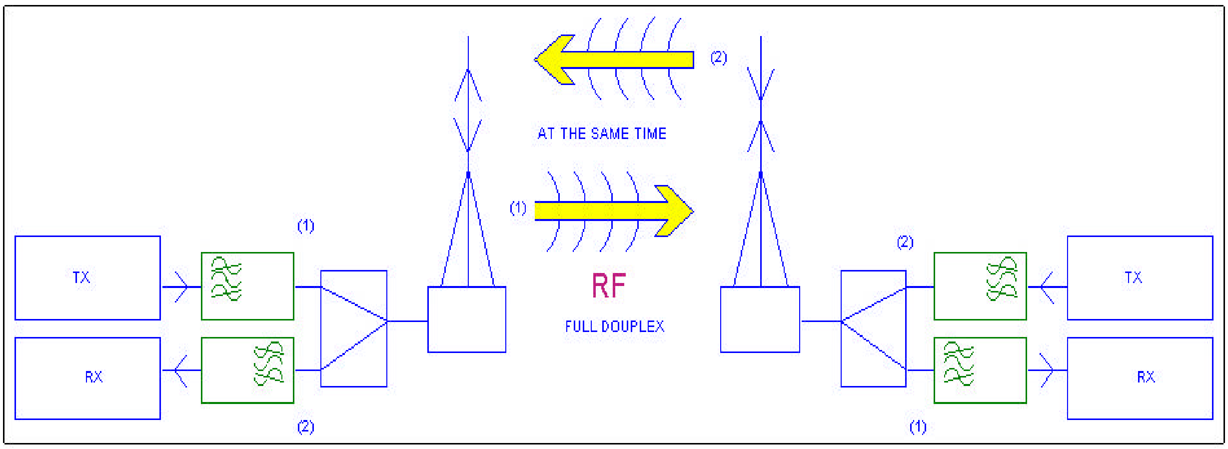
Simplex



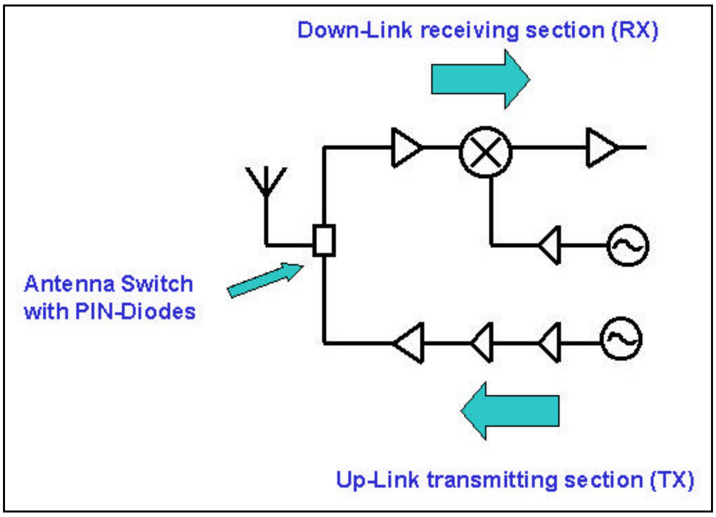
Half duplex

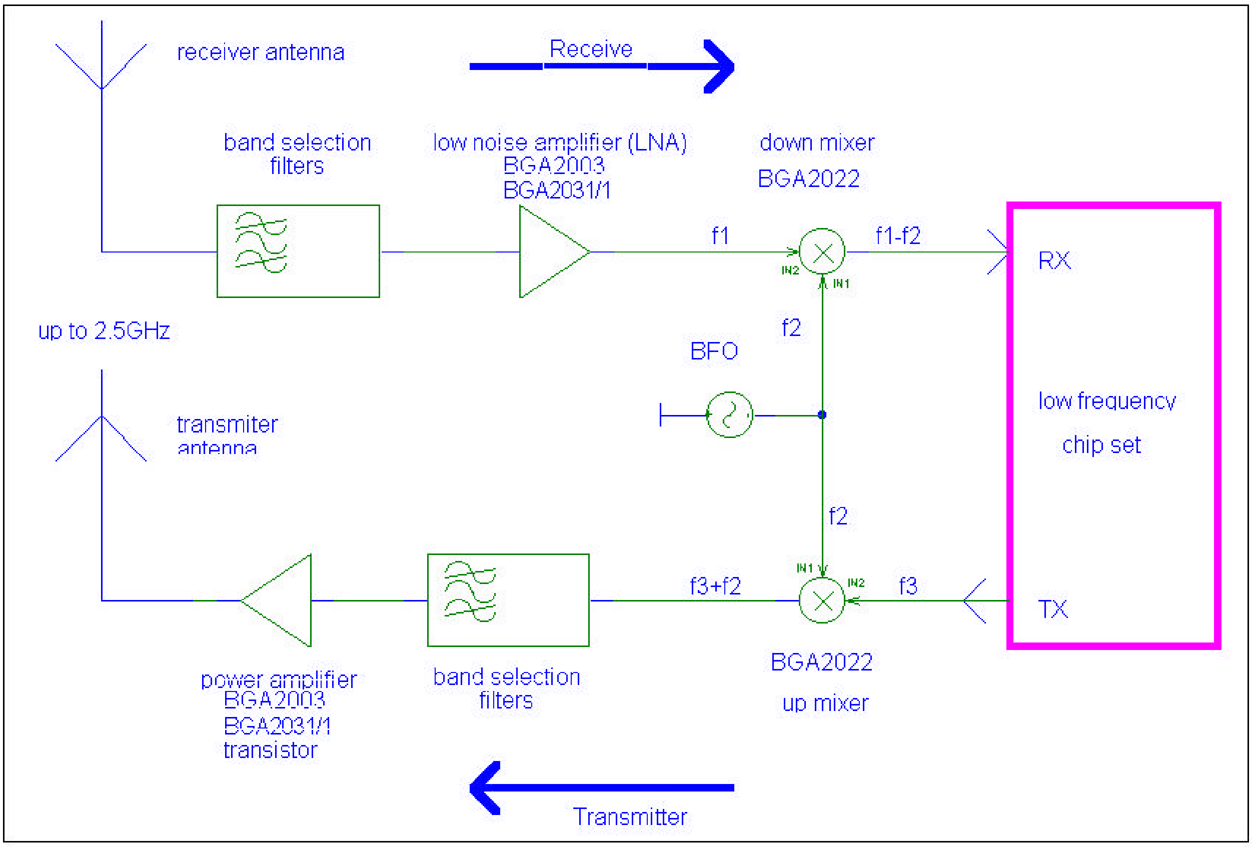


Full duplex



1.3 RF Front-End

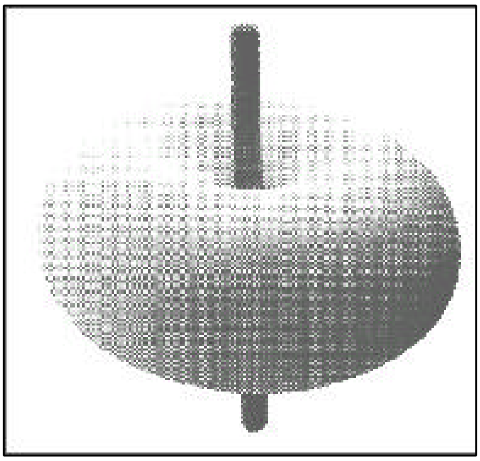




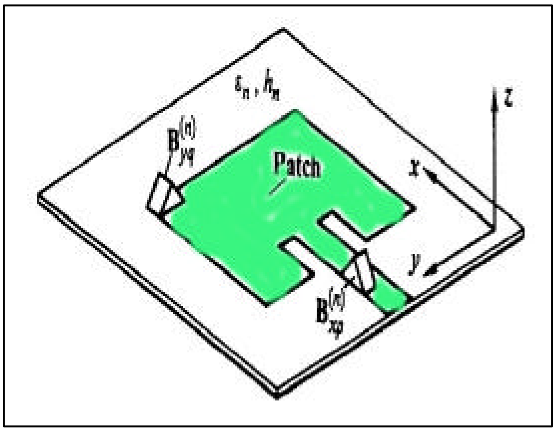
3.4 Function of an antenna

In standard application the RF output signal of a transmitter power amplifier is transported via a coaxial cable to a suitable location where the antenna is installed. Typically the coaxial cable has an impedance of 50Ω (75Ω for TV/Radio). The ether, that is the room between the antenna and infinite space, also has an impedance value. This ether is the transport medium for the traveling wireless RF waves from the transmitter antenna to the receiver antenna. For optimum power transfer from the end of the coaxial cable (e.g. 50Ω) into the ether (theoretical $Z=120\cdot\pi\cdot\Omega=377\Omega$), we need a “power matching” unit. This matching unit is the antenna. It does match the cable’s impedance to the space’s impedance. Depending on the frequency and specific application needs there are a lot of antenna configurations and construction variations available. The simplest one is the isotropic ball radiator, which is a theoretical model used as a mathematical reference.

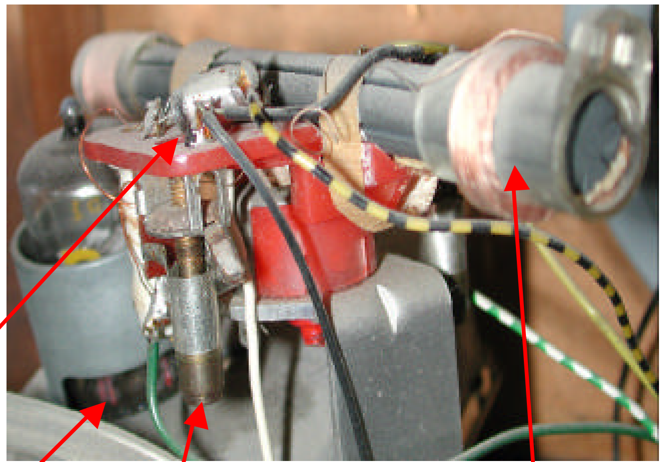
The next simplest configuration and a practical antenna in wide use is the dipole, also called the dipole radiator. It consists of two axial arranged sticks (Radiator). Removal of one Radiator results in to the “vertical monopole” antenna, as illustrated in the adjacent picture. The vertical monopole has a “donut-shaped” field centered on the radiating element.

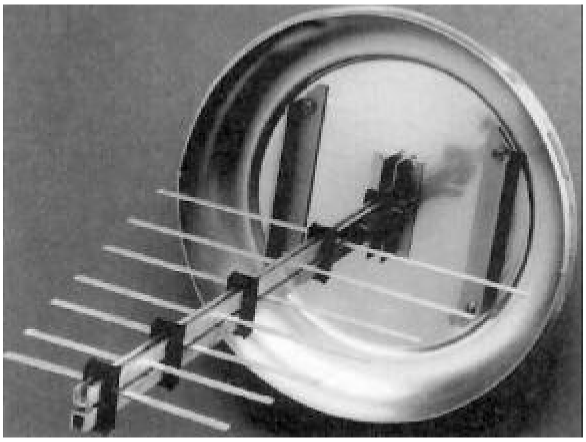


Higher levels of integration of the circuitry and reductions in cost also influence antenna design. Based on the EM field radiation of Strip lines made by printed circuit boards, a PCB antenna structures were developed called a “Patch”-Antenna as illustrated in the adjacent picture. Use of ceramic instead of epoxy dielectric do again shrink mechanical dimensions.



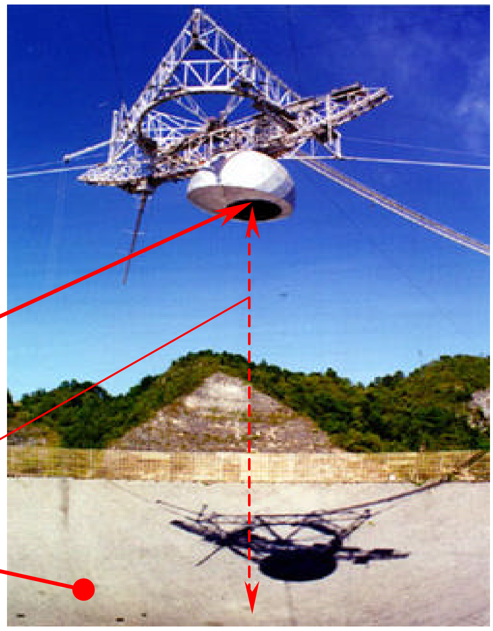
In the application range of LF-MF-HF their was used Ferrite Rod Antennas as illustrated in the adjacent picture. It do compress the magnetic fields into the Ferrite core. This appears like an amplifier for magnetic RF fields. The coils do pick up like a transformer. They are a part of the pre-selection LC tank for image rejection and channel selection. This tuner is a part of an at least 40yr's old Nordmende Elektra vacuum tube radio (still working at the author). For illustration of the dimensions, a Monolithic Microwave IC is placed in front of a solder point.





Logarithmic Periodic Antenna for 406-512MHz

UHF Broadband Discone Antenna

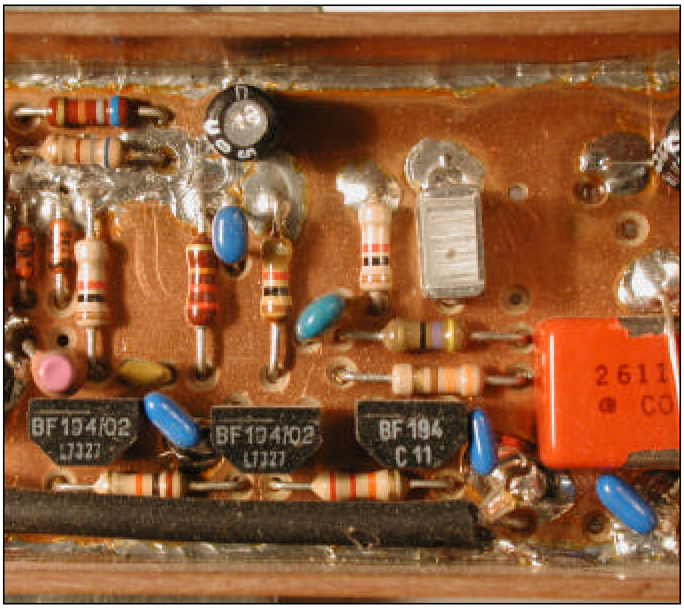


900t's have Fed point of a L-Band Microwave antenna + 50MHz-10GHz Antennas located 137m above the dish antenna center (Radio-Telescope Arecibo, Puerto Rico) The dish has a diameter of 305m and a depth of 51m for the SETI@home receiver. In the focus is located the receiver. The receiver is cold down to 50k by the use of Fluid Helium for low noise operation. That's need for searching for signals transmitted from extraterrestrial intelligence. Response is possible by a balanced Klystron amplifier with 2.5kW output peak power. (120KV/4.4A power supply)

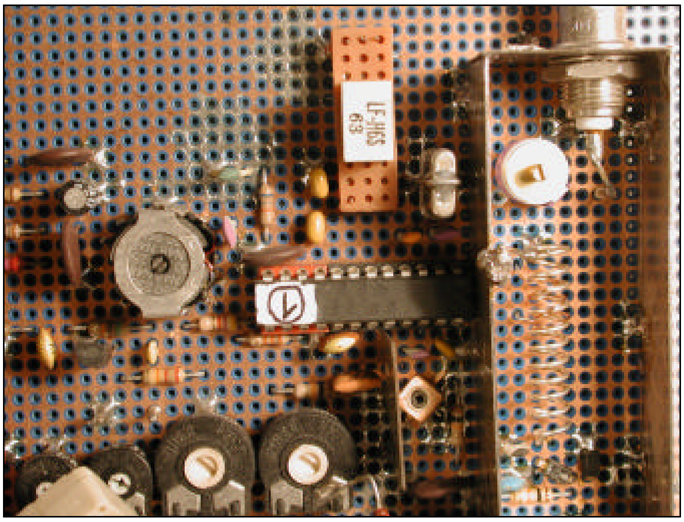
1.5 Examples of PCB design

- Low frequency design (up to several tens of MHz)
- RF design (tens of MHz to several hundreds of MHz)
- Microwave design (GHz range)

1.5.1 Prototyping

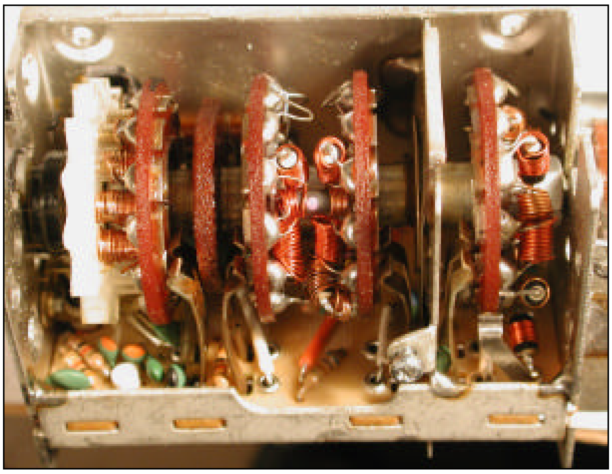


HF-Range:
(Prototype) Top side GND, back side manual wires forms a 3 stage short wave antenna amplifier.

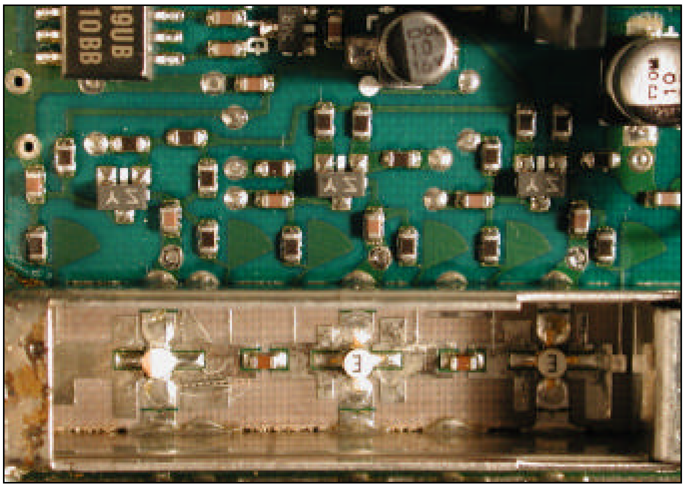


HF to VHF-Range:
(Prototype) Receiver Front-End:
Top side GND, back side manual wires forms a 144MHz double Superhet receiver with 10.7MHz + 455KHz IF.

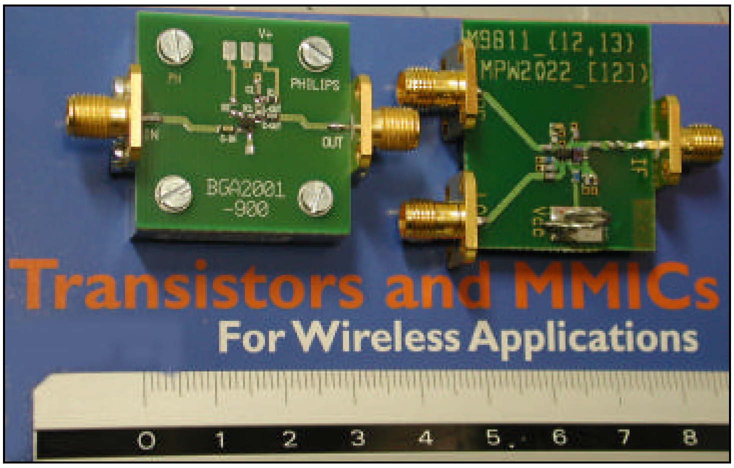
1.5.2 Final PCB



VHF/UHF-Range:
TV-Tuner: PCP and flying parts on the switch (history); some times prototyping technology at RF



UHF/SHF-Range:
Sat Microwave Front-End in Microstrip Technology



VLF to SHF-Range:
Demoboards BGA2001 and BGA2022 from Philips Semiconductors in Microstrip Technology

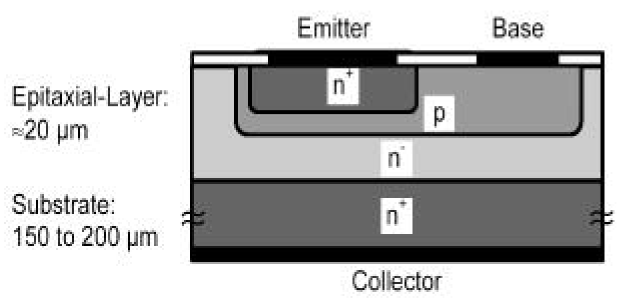
1.6 Transistor Semiconductor Process

1.6.1 General-Purpose Small-signal bipolar

The transistor is built up from three different layers:

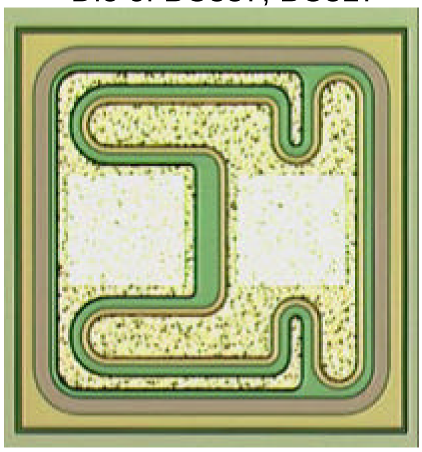
- Highly doped emitter layer
- Medium doped base area
- Low doped collector area.

NPN Transistor cross section



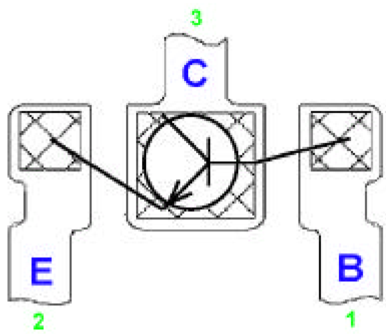
The highly doped substrate serves as carrier and conductor only.

Die of BC337, BC817



During the assembly process the transistor die is attached on a lead frame by means of gluing or eutectic soldering. The emitter and base contacts are connected to the lead frame (leads) through (e.g. Gold, Aluminium, …) bond wires in e.g. an ultrasonic welding process.

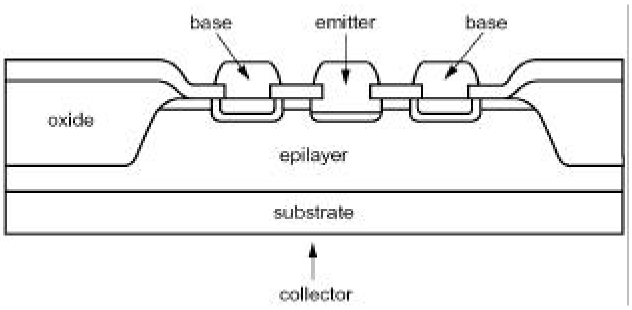
SOT23 standard lead frame



1.6.2 Double Polysilicon

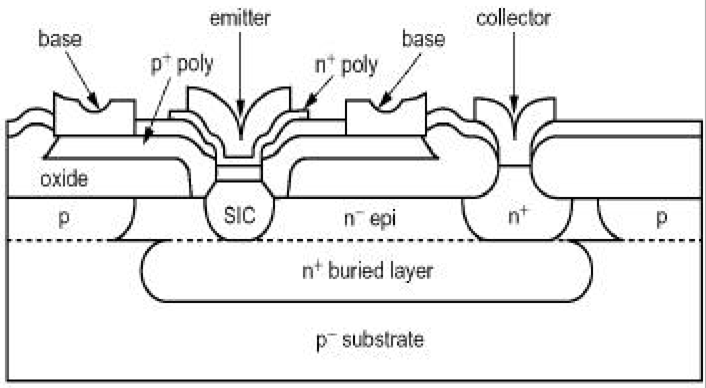
For the latest Silicon-based bipolar transistors and MMICs, Philips has developed a Double Polysilicon process to achieve excellent performance.

The mobile communications market and the use of ever-higher frequencies have do need of low-voltage, high-performance, RF wideband transistors, amplifier modules and MMICs. The “double-poly” diffusion process makes use of an advanced, transistor technology that is vastly superior to existing bipolar technologies.



Existing advanced bipolar transistor

With double poly, a polysilicon layer is used to diffuse and connect the emitter while another polysilicon layer is used to contact the base region. Via a buried layer, the collector is brought out on the top of the die.
As with the standard transistor, the collector is picked up via the backside substrate and attachment to the lead frame.



- **Advantages of double-poly-Si RF process:**
- Higher frequencies (>23GHz)
- Higher power gain G_{max}, e.g., 22dB/2GHz
- Lower noise operation
- Higher reverse isolation
- Simpler matching
- Lower current consumption
- Optimized for low supply voltages
- High efficiency
- High linearity
- Better heat dissipation
- Higher integration for MMICs (***SSI***= **S**mall-**S**cale-***I***ntegration)

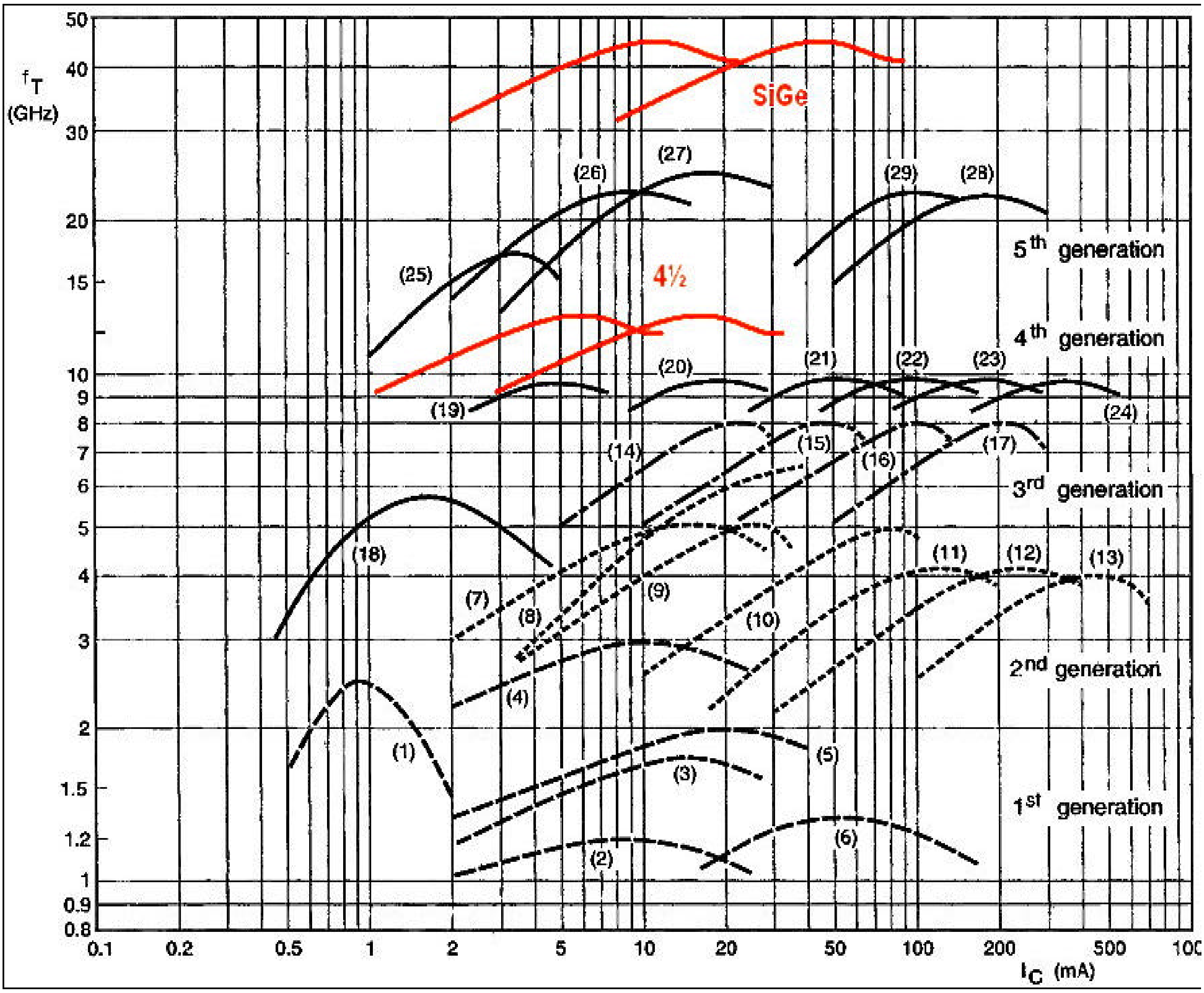
- **Applications**

Cellular and cordless markets, low-noise amplifiers, mixers and power amplifier circuits operating at 1.8 GHz and higher), high-performance RF front-ends, pagers and satellite TV tuners.

- **Typical vehicles manufactured in double-poly-Si:**

▪ MMIC Family:	BGA20xy, and BGA27xy
▪ 5th generation wideband transistors:	BFG403W/410W/425W/480W
▪ RF power amplifier modules:	BGY240S/241/212/280

1.6.3 RF Bipolar Transistor & MMIC Performance overview



Appendix C: RF Design-basics

1.1 RF Fundamentals

1.1.1 Frequency and time domain

1.1.1.1 Frequency domain operations

Typical vehicles-effects and test-equipment:

- Metallic sound and distortions of a low-cost PC loudspeaker
- Audio analyzer (measuring the quality of the audio signal, like noise and distortion)
- FFT Spectrum analyzer (in the medium frequency range from a few Hertz to several MHz)
- Modulation analyzer (investigation of RF modulation e.g., AM, FSK, GFSK, et. al.)
- Spectrum analyzer (display the signal's spectral quality, e.g., noise, intermodulation, gain)

The mathematical Fourier Transform algorithm analyses the performance of a periodical time depending signal into the frequency domain. For a one-shot signal the Fourier Integral Transformation is used. On the bench, test issues are over-taken by the spectrum analyzer or by a ***FFT*** analyzer (***F***ast ***F***ourier ***T***ransformation). With the spectrum analyzer the frequency spectrum of the device under test (***DUT***) are scanned into bands (e.g., by tuned filters) and measured in a detector (like a periodic tuned radio with displaying of the field strength). The FFT analyzer is essentially a computer capable of performing a ***DSP*** (***D***igital ***S***ignal ***P***rocessor) function. This DSP has a built-in hardware-based circuit for very fast solution of algorithmic problems like the ***DFFT*** (***D***iscrete ***F***ast ***F***ourier ***T***ransformation). This DFFT algorithmic can calculate the frequency spectrum of an incoming signal. DSP processors are used in today's mobile equipment to provide base band or IF signal processing, sound cards for computers, industrial machinery, communication receivers, motor control, and other complex signal processing functions.

In RF and microwave applications, the frequency domain is very important for measurement techniques, because oscilloscopes cannot display extremely high frequency signals and have probe impedances causing excessive load and detuning by their input capacitance. A spectrum analyzer has much better sensitivity, a much larger dynamic range capability and a broadband 50Ω/75Ω input.

Example: An oscilloscope can simultaneously display signals with a voltage ratio of 10 to 20 between the smallest and largest signals (a dynamic range ~20dB). RF spectrum analyzers can display power signal (levels) with a ratio between the largest signal and the smallest signal of more than 10^6 at the same time on the display (dynamic range >>60dB). Intermediate frequency (IF) amplifiers of typical receivers have gains of 40 to 60dB, meaning the amplifier output signal can be 10^4 to 10^6 larger than the input signal. The spectrum analyzer can display both input and output signals simultaneously with good accuracy on to the logarithmic display for both. On an oscilloscope (with a linear display) setting the amplitude of the output signal at full-scale allows you to perhaps see what appears to be some noise ripple on the axis for the input signal. Typical modern oscilloscopes support frequency ranges up to few GHz. Modern spectrum analyzers start at several tenths of kHz and go up to several tens of GHz. Special function spectrum analyzers provide signal analysis more than 100GHz.

1.1.1.2 Time domain operations

Typical bench vehicle and applications:

- Booting beeps in the PC computer's loudspeaker
- The oscilloscope (displays the signal's action over the time)
- The RF generator (generates very clean sine wave test signals with various modulation options)
- The ***T***ime ***D***omain ***R***eflectometry analyzer (***TDR***) (e.g., analyzing cable discontinuities)
- Jitter in clock-recovery circuits
- Eye diagrams

In the time domain the variation of the amplitude is displayed versus the time on a screen. Very low speed activities such as temperature drift versus aging of an oscillator or seismic activity are printed by special plotters in real-time on paper. Faster actions are better displayed by oscilloscopes. Signals can be stored on the oscilloscope screen by the use of storage tubes (history), or by the use of built-in digital storage (RAM). In the time domain, phase differences between different sources or time-dependent activities can be analyzed, characterized or modified.

In RF applications displays show demodulation actions, base-band signals or control functions of a CPU. The advantage of the oscilloscope is the high resistive impedance of the probes. It's disadvantage is the input capacity of several pico Farads (pF) causing high frequency AC loading of the circuit, which affects both the measured RF circuit and distorts the measurement data presented.

Mixers are inherently non-linear devices because their chief function is multiplication of signals. On the input side the RF signal must be treated linearly. Mixer ***3rd order intercept point*** (***IP3***) performance characterizes the quality of handling the RF signals and the amount non-linearity introduced.

Example illustrating an application circuit in the frequency domain and in the time domain:

Issue: Receiving the commercial radio broadcasting program SWR3 in the short-wave 49m band from the German transmitter-Mühlacker on 6030 kHz. This transmitter has an output power of 20000W. Design the mixer using a 455 kHz IF amplifier.
Reference: http://www.swr.de/frequenzen/kurzwelle.html

System design of the ***local oscillator***: LO = RF + IF = 6030 kHz + 455 kHz = 6485 kHz
The ***image frequency*** is found at IRF = LO + IF = 6485 kHz + 455 kHz =6913 kHz
Optimum mixer operation is medium gain for IF and RF and damping of RF and LO transfer to the IF port (isolation). As an example, we choose the ***BFR92***. This transistor can also be used for much higher frequency mixer applications like FM radios, televisions, ISM433, and other applications.

As shown in the formulas above, the ***R***adio ***F***requency (***RF***) signal is mixed with the ***L***ocal ***O***scillator (***LO***) to generate the ***I***ntermediate ***F***requency (***IF***) output products.

To improve the mixer gain, several part values were varied. This circuit is a theoretical example for discussion purposes only. Further optimization should be done by investigation on bench. In the example the input signal sources V6 and V7 are series connected. In the reality this can be done by e.g. A transformer. The simulation was done under PSpice with the following setup: Print Step=0.1ns; Final Time=250µs; Step Ceiling=1ns. This long simulation length and fine resolution is necessary for useful results in the frequency spectrum analysis down to 400KHz.

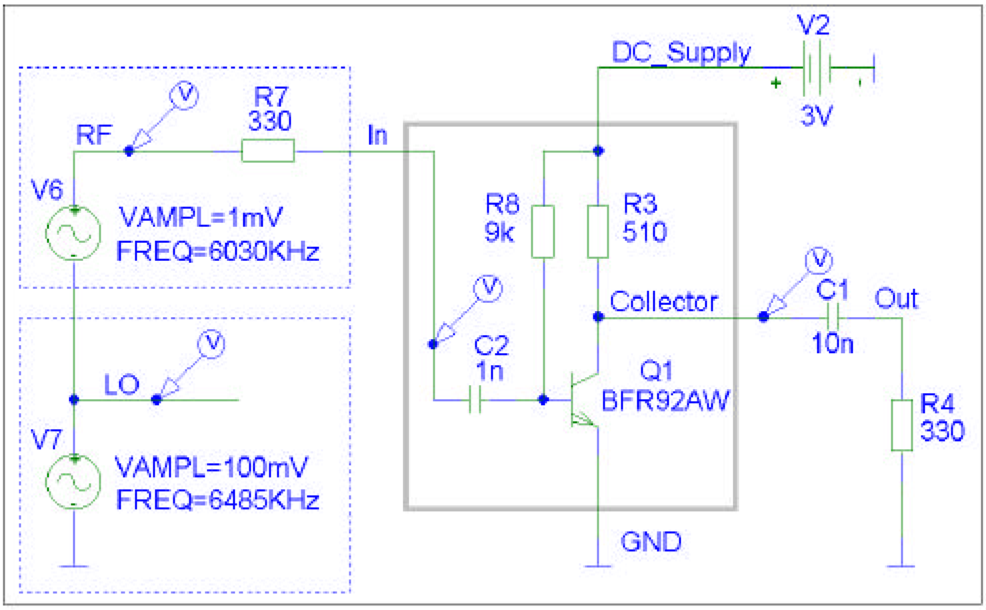


Figure 1: Final mixer circuit without output IF tank

Varying of R8 shows the influence of the mixer gain at the 455 kHz output frequency.

R8	***6k***	***7k***	***8k***	***9k***	***10k***	***15k***	***20k***	***25k***
455KHz	0.32mV	2.21mV	3.37mV	3.66mV	3.62mV	2.33mV	1.43mV	1.44mV
12515KHz	0.29mV	2mV	2.94mV	3.11mV	2.97mV	1.52mV	0.83mV	0.5mV

From the experiments we chose R8 = 9 kΩ for best output amplitude.

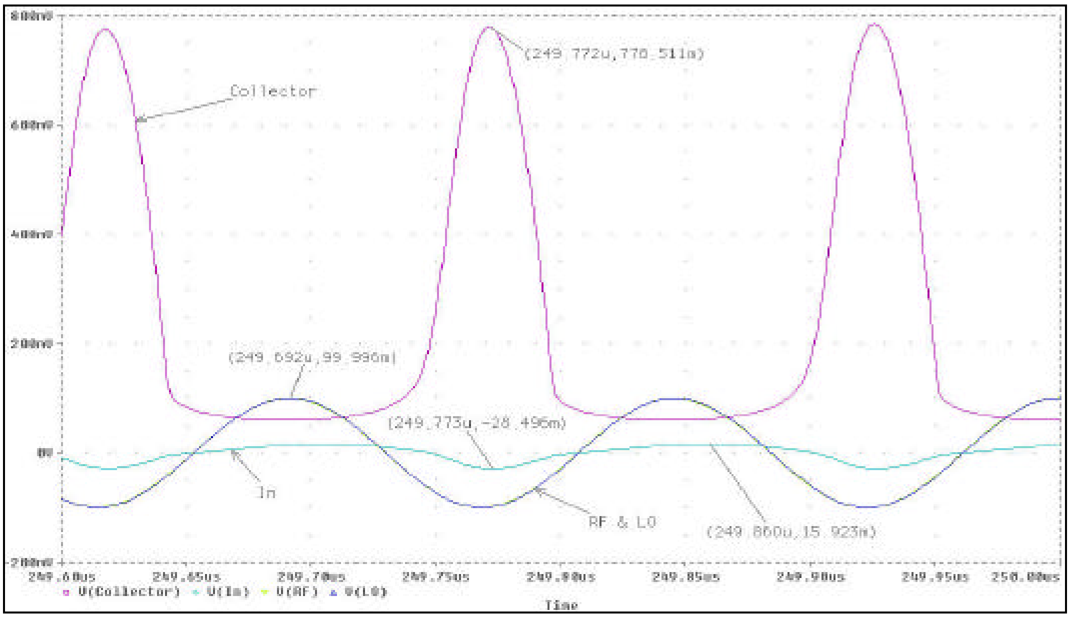


Figure 2: The mixer in the time domain arena

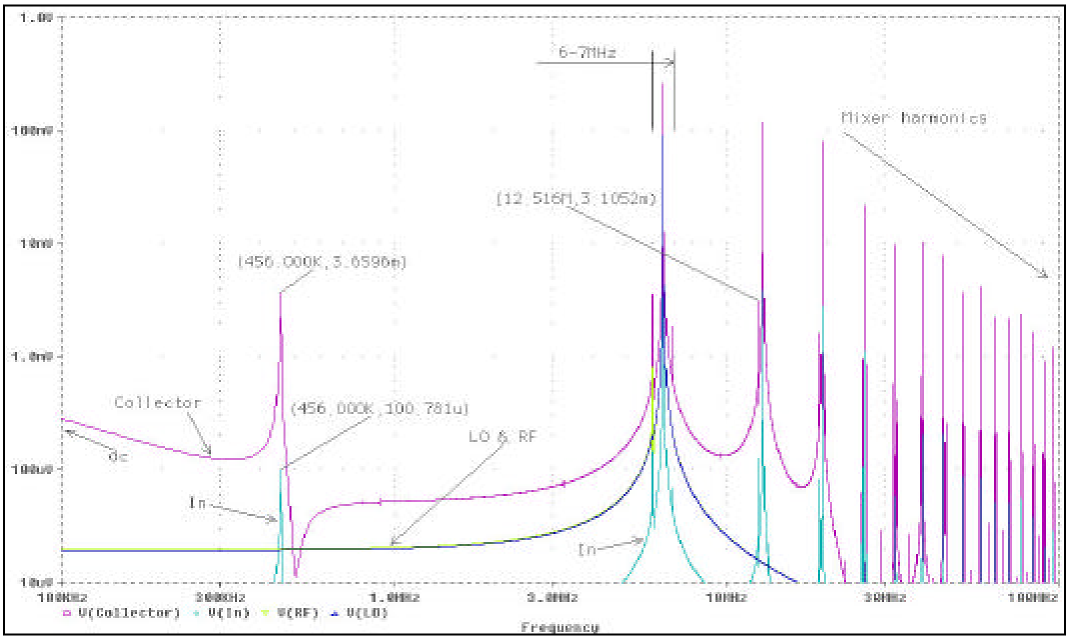


Figure 3: The mixer in the frequency domain arena

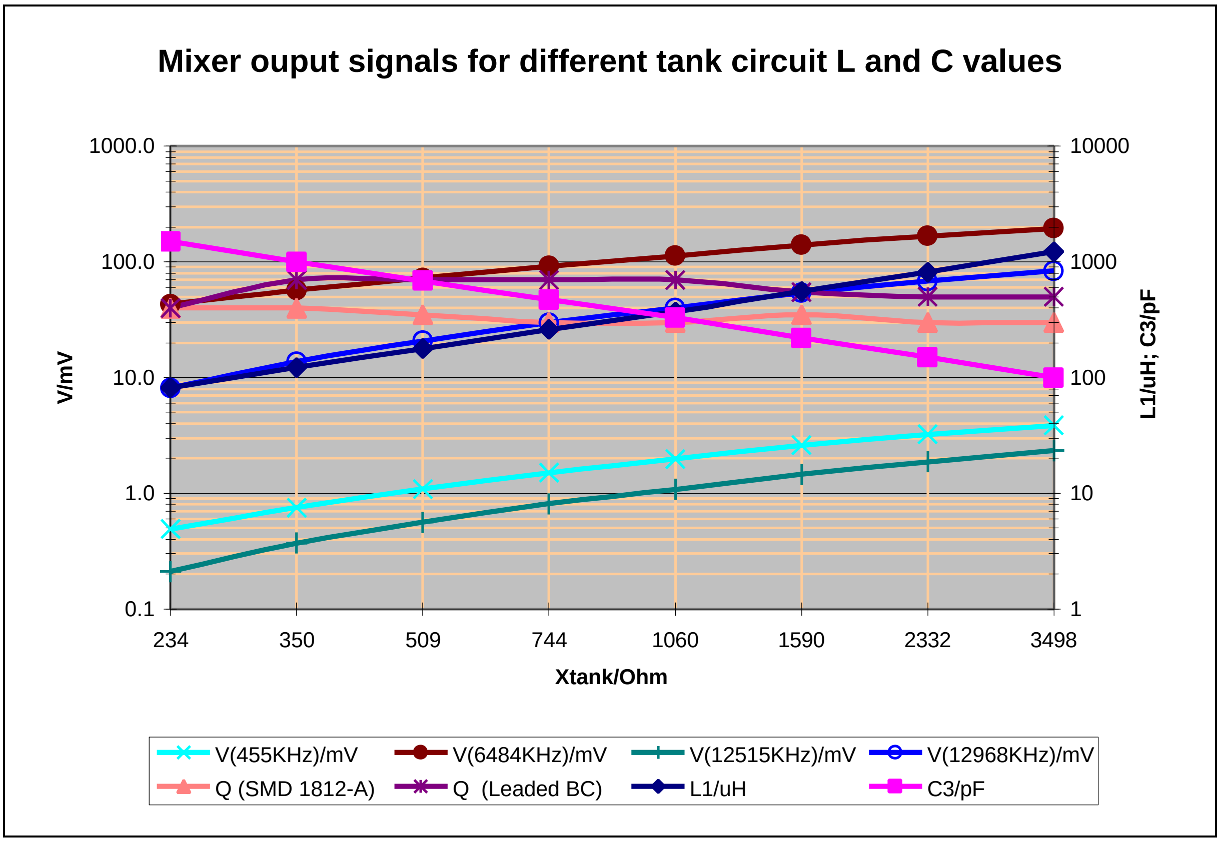


Figure 4: Mixer output voltage versus the tank circuit's characteristic resonance impedance

Further investigated must be the available IF bandwidth. A narrow IF bandwidth reduces the fidelity of the demodulated signal but improves noise related issues and selectivity of a receiver.

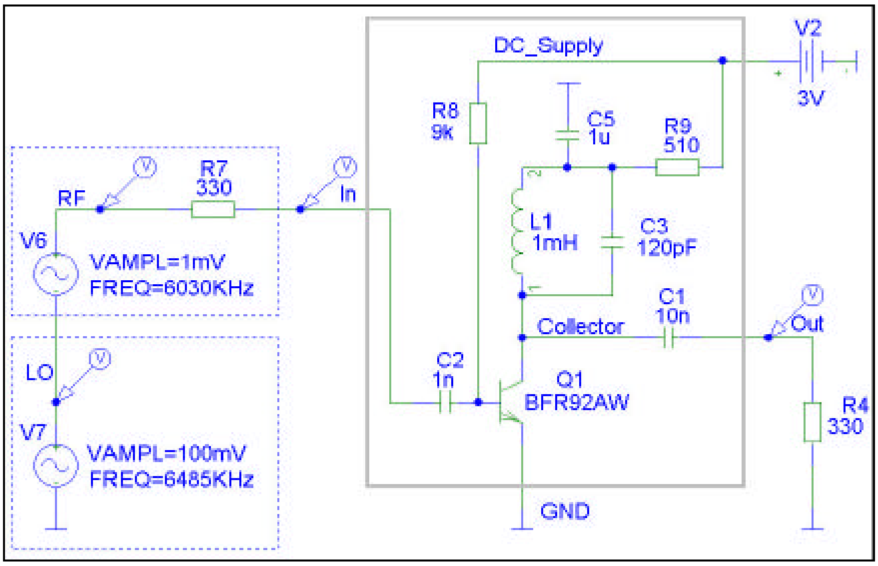


Figure 5: The mixer with an IF tank circuit

This chapter illustrated a mixer operation in both time and frequency domains. Illustrated was circuit design by "trial and error" coupled with the use of a CAD program with a lot of simulation time. A better approach would be the use of a design strategy and calculation of the exact required values and then final CAD optimization. The devices must be accurately specified (S-parameters) and models (e.g., 2-port linear model network) must be available for computer simulation. The use of time domain simulators with different algorithms (eg. harmonic balance) accelerates the simulation. Philips Semiconductors offers s-parameters for small signal discrete devices. Because optimum power transfer is important in RF application, we must think about the quality of inter-stage circuit matching, qualified by the reflection coefficient. This will be handled in the next chapters. Please note that Philips Semiconductors offers a ***M**onolithic **M**icrowave **I**ntegrated **C**ircuit* (***MMIC***) mixer, a ***BGA2022***, with a 50W input impedance. This device has built-in biasing circuit and offer excellent gain and linearity.

1.1.2 RF waves

RF ***electro-magnetic (EM)*** signals travel outward like ***waves*** in a pond that has a stone dropped into it. The EM waves are governed by the laws that particularly apply to optical signals. In a homogeneous vacuum without external influences EM waves travel at a ***speed of C_o=299792458 m/s***. Travelling in substrates, wires, or within a non-air ***dielectric*** material put into the travelling path slows the speed of the waves proportional to the root of the dielectric constant:

$$\boxed{v = \frac{C_O}{\sqrt{e_{reff}}}}$$

ε_{reff} is the ***substrate's dielectric constant***.

With "v" we can calculate the ***wavelength***, as: $\boxed{l = \frac{v}{f}}$

Example1: Calculate the speed of an electromagnetic wave in a ***P**rinted **C**ircuit **B**oard* (***PCB***) manufactured using a FR4 epoxy material and in a metal-dielectric-semiconductor capacitor of an integrated circuit.

Calculation: In a metal-dielectric-semiconductor capacitor the dielectric material can be Silicon-Dioxide (SiO_2) or Silicon-Nitride (Si_3N_4).

$$v = \frac{C_O}{\sqrt{e_{reff}}} = \frac{299792458 m/s}{\sqrt{4.6}} = 139.78 \cdot 10^6 m/s$$

FR4	➔	ε_{reff}=4.6	➔	v=139.8•10^6m/s
SiO_2	➔	ε_{reff}=2.7 to 4.2	➔	v=182.4•10^6m/s to 139.8•10^6m/s
Si_3N_4	➔	ε_{reff}=3.5 to 9	➔	v=160.4•10^6m/s to 99.9•10^6m/s

Example2: What is the wavelength transmitted from the commercial SW radio broadcasting program SWR3 in the 49 meter (m) band on 6030 kHz in air, and within a FR4 PCB?

Calculation: The ε_{reff} of air is close to vacuum. ➔ $\varepsilon_{reff} \approx 1$ ➔ $n = c_O$

Wavelength in air: $l_{air} = \frac{C_O}{f} = \frac{299792458 m/s}{6030 KHz} = 49.72 m$

From Example 1 we take the FR4 dielectric constant to be ε_{reff} = 4.6, then n=139.8•10^6m/s and calculate the wavelength in the PCB as: λ_{FR4} = 23.18 meters

A forward-traveling wave is transmitted (or injected) by the source into the traveling medium (whether it be the ether, a ***substrate***, a ***dielectric***, wire, ***Microstrip***, ***wave-guide*** or other medium) and travels to the load at the opposite end of the medium. At junctions between two different dielectric materials, a part of the forward-traveling wave is reflected back towards the source. The remaining part continues traveling towards the load.

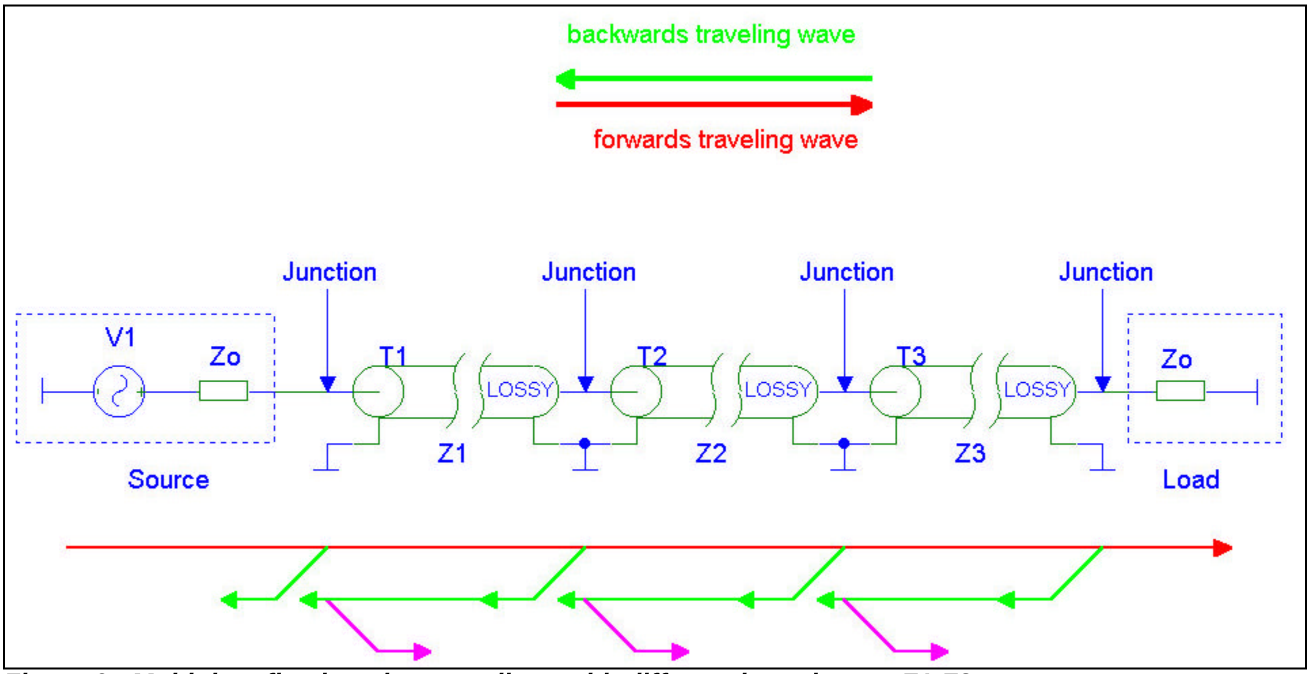
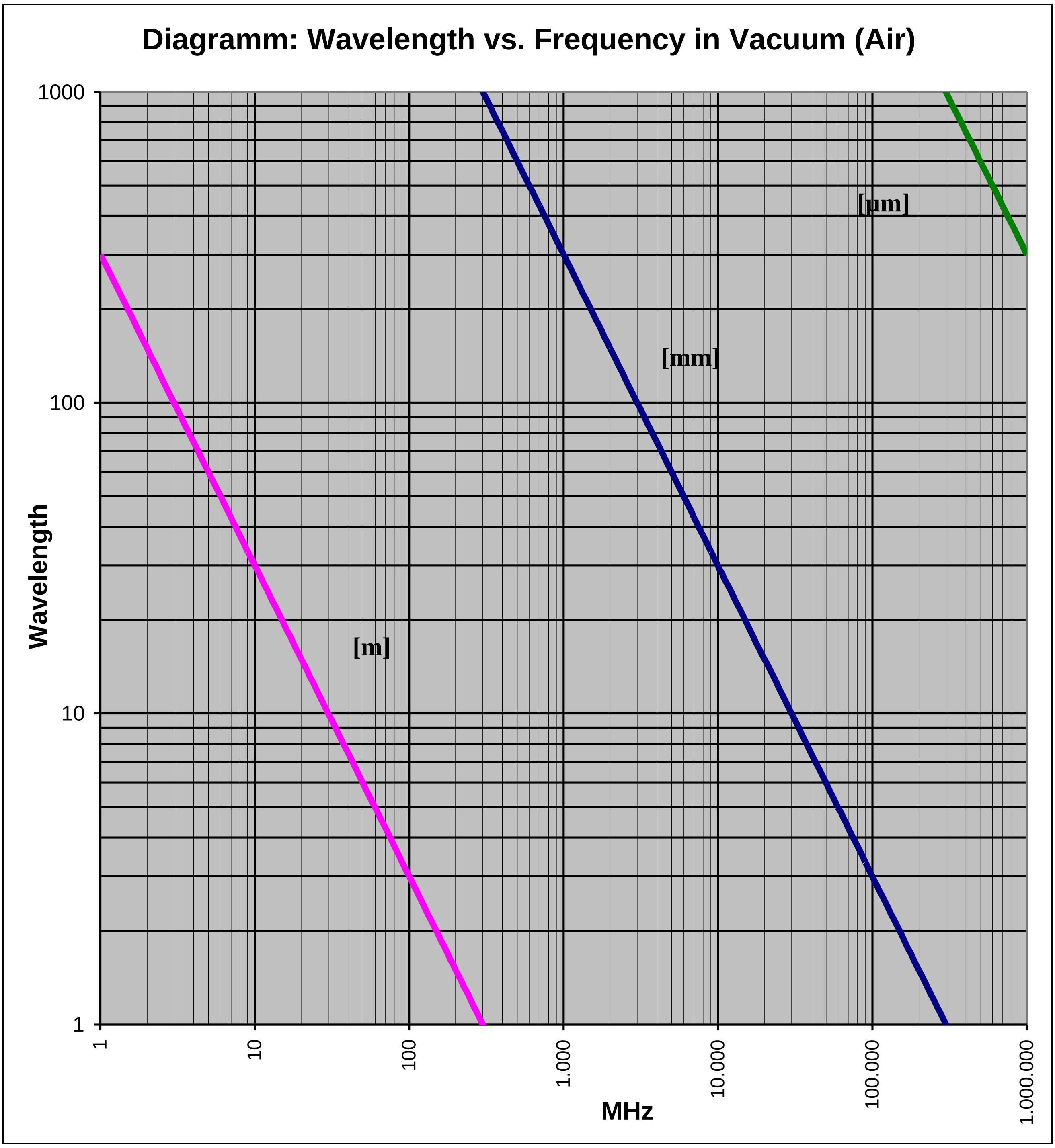


Figure 6: Multiple reflections between lines with different impedances Z1-Z3

In Figure 6 ***reflections*** of the forward-traveling main wave (red) are caused between materials with different impedance values (Z1, Z2, Z3). As shown, a backward-reflected wave (green) can be again reflected into a forward-traveling wave in the direction towards the load (shown as violet in Figure 6). In the case of optimum ***matching*** between different dielectric mediums, no signal reflection will occur and maximum power is forwarded. The amount of reflection caused by junctions of lines with different impedances, or line ***discontinuities***, is determined by the ***reflection coefficient***. This is explained in the next chapter.



Example: Select your frequency (ISM433) crossing a trace (blue) you can read the wavelength (70cm)

1.1.3 The Reflection Coefficient

As discussed previously a forward-traveling wave is partially reflected back at junctions with line impedance discontinuities, or mismatches. Only the portion of the forward traveling wave (arriving at the load) will be absorbed and processed by the load. Because of the frequency-dependent speed of the propagating waves in a dielectric medium, there will be a delay in the arrival of the wave at the load point over what a wave traveling in free space would have (phase shift). Mathematically this behavior is modeled with a vector in the complex Gaussian space. At each location of the travel medium (or wire), wave-fronts with different amplitude and phase delay are heterodyned. The resulting energy envelope of the waves along the wire appears as ripple with maximum and minimum values. The phase difference between maximums does has the same value as the phase difference between minimums. This distance is termed the ***half-wavelength, or l/2 (also termed the normalized phase shift of 180°)***.

Example: A line with mismatched ends driven from a source will have standing waves. These will result in minimum and maximum signal amplitudes at defined locations along the line. Determine the approximate distance between worst-case voltage points for a ***Bluetooth*** signal processed in a printed circuit on a FR4 based substrate.

Calculation: Assumed speed in FR4: v=139.8·10^{6}m/s

Wavelength: $$l_{air} = \frac{v_{FR4}}{f_{BT}} = \frac{139.78 \cdot 10^6\, m/s}{2.4GHz} = 58.24mm$$

The distance minimum to maximum is called the ***quarter wavelength, or l/4 (also termed the normalized phase shift of 90°)***.

Min-Max distance in FR4: $$l/4 = \frac{58.24mm}{4} = 14.56mm$$

- At the minimum we have minimum voltage, but maximum current.
- At the maximum we have maximum voltage, but minimum current.
- The distance between a minimum and a maximum voltage (or current) point is equal to *l/4*.

The reflection coefficient is defined by the ratio between the backward-traveling voltage wave and the forward-traveling voltage wave:

Reflection coefficient: $$\boxed{r_{(x)} = \frac{U_{b(x)}}{U_{f(x)}}}$$

Reflection loss or ***return loss***: $$\boxed{r_{dB} = 20dB \cdot \log\left|r_{(x)}\right| = 20dB\left\{\log\left|U_{b(x)}\right| - \log\left|U_{f(x)}\right|\right\}}$$

The index "(x)" indicates different reflection coefficients along the line. This is caused by the distribution of the standing wave along the line. The return loss indicates, in dB, how much of the wave is reflected, compared to the forward-traveling wave.

Often the input reflection performance of a 50W RF device is specified by the **Voltage Standing Wave Ratio** (***VSWR***), also called the ***SWR***.

VSWR: $s = SWR = VSWR = \frac{U_{max}}{U_{min}}$ Matching factor: $m = \frac{1}{s}$ which for practical applications requires the VSWR>1.

Some typical values of the VSWR:
100% mismatch caused by an open or shorted line: r = 1 and VSWR = ¥
Optimum (theoretical) matched line: r = 0 and VSWR = 1
In all practical situations “r” varies between 0 < r < 1 and 1 < VSWR < ¥

Calculating the reflection factor: $r = |r_{(x)}| = \frac{SWR-1}{SWR+1}$

Using some mathematical manipulation: $r = \frac{\frac{U_{max}}{U_{min}} - 1}{\frac{U_{max}}{U_{min}} + 1}$ results in: $r = \frac{U_{max} - U_{min}}{U_{max} + U_{min}}$

Reflection coefficients of a certain impedances (eg. a load) leads to: $r = \frac{Z - Z_O}{Z + Z_O}$

with Z_o = nominal system impedance (50Ω, 75Ω)

As explained, the standing waves cause different amplitudes of voltage and current along the wire. The ratio of these two parameters is the impedance $Z_{(x)} = \frac{V_{(x)}}{I_{(x)}}$ at each locations, (*x*). This means a line with length (*L*) and a mismatched load $Z_{(x = L)}$ at the wire end location (*x=L*) will show at the sources location (*x=0*) a wire length dependent impedance’s $Z_{(x=0)\ f(\ell)} = \frac{V_{(x=0)}}{I_{(x=0)}}$.

Example: There are several special cases (tricks), which can be used in microwave designs. Mathematically it can be shown that a wire with the length of $\ell = \frac{l}{4}$ and an impedance Z_L will be a ***quarter wavelength transformer***:

$l/4$ - impedance transformer: $Z_{(x=\ell)} = \frac{Z_L^{\ 2}}{Z_{(x=0)}}$

This can be used in SPDT based *p-i-n* diode switches or in DC bias circuits because an RF short (like a large capacitor) is transformed into infinite impedance with low resistive dc path (under ideal conditions).

As indicated in Figure 6, and shown by the RF traveling-wave basic rules, the performances of matching, reflection and individual wire performances affect bench measurement results, caused by impedance transformation along the wire. Due to this constraint, each measurement set-up must be calibrated by precision references.

Examples of RF calibration references are:
- Open - Through
- Short - Sliding Load
-Match

The set-up calibration tools can undo unintended wire transformations, discontinuities from connectors, and similar measurement intrusion issues. This prevents ***D***evice ***U***nder ***T***est (***DUT***) measurement parameters from being affected with mechanical bench set-up configurations.

Example: a) Determine the input VSWR of ***BGA2711*** MMIC wideband amplifier for 2GHz, based on data sheet characteristics.
b) What kind of resistive impedance(s) can theoretically cause this VSWR?
c) What is the input return loss measured on a 50Ω coaxial cable in a distance of λ/4?

Calculation: BGA2711 at 2 GHz: r_{IN} = 10dB

$$r = \frac{SWR-1}{SWR+1} \rightarrow r \cdot SWR + r = SWR - 1 \rightarrow \boxed{SWR = \frac{1+r}{1-r}} \quad r = 10^{\frac{-r_{dB}}{20dB}} = 10^{\frac{-10dB}{20dB}} = 0.3162$$

$$\rightarrow SWR_{IN} = \frac{1+0.3162}{1-0.3162} = 1.92 \quad r = \frac{Z-Z_O}{Z+Z_O} \rightarrow Z - r \cdot Z = r \cdot Z_O + Z_O \rightarrow \boxed{Z = Z_O \frac{1+r}{1-r}}$$

Comparison: $Z = Z_O \frac{1+r}{1-r}$ & $SWR = \frac{1+r}{1-r}$ → $Z = Z_O \cdot SWR$

We know only the magnitude of (*r*) but not it's angle. By definition, the VSWR must be larger than 1. We then get two possible solutions:

$\boxed{SWR_1 = \frac{Z_{max}}{Z_O}}$ and $\boxed{SWR_2 = \frac{Z_O}{Z_{min}}}$ Z_{max}=1.92∗50Ω=96.25Ω; Z_{min}=50Ω/1.92=25.97Ω

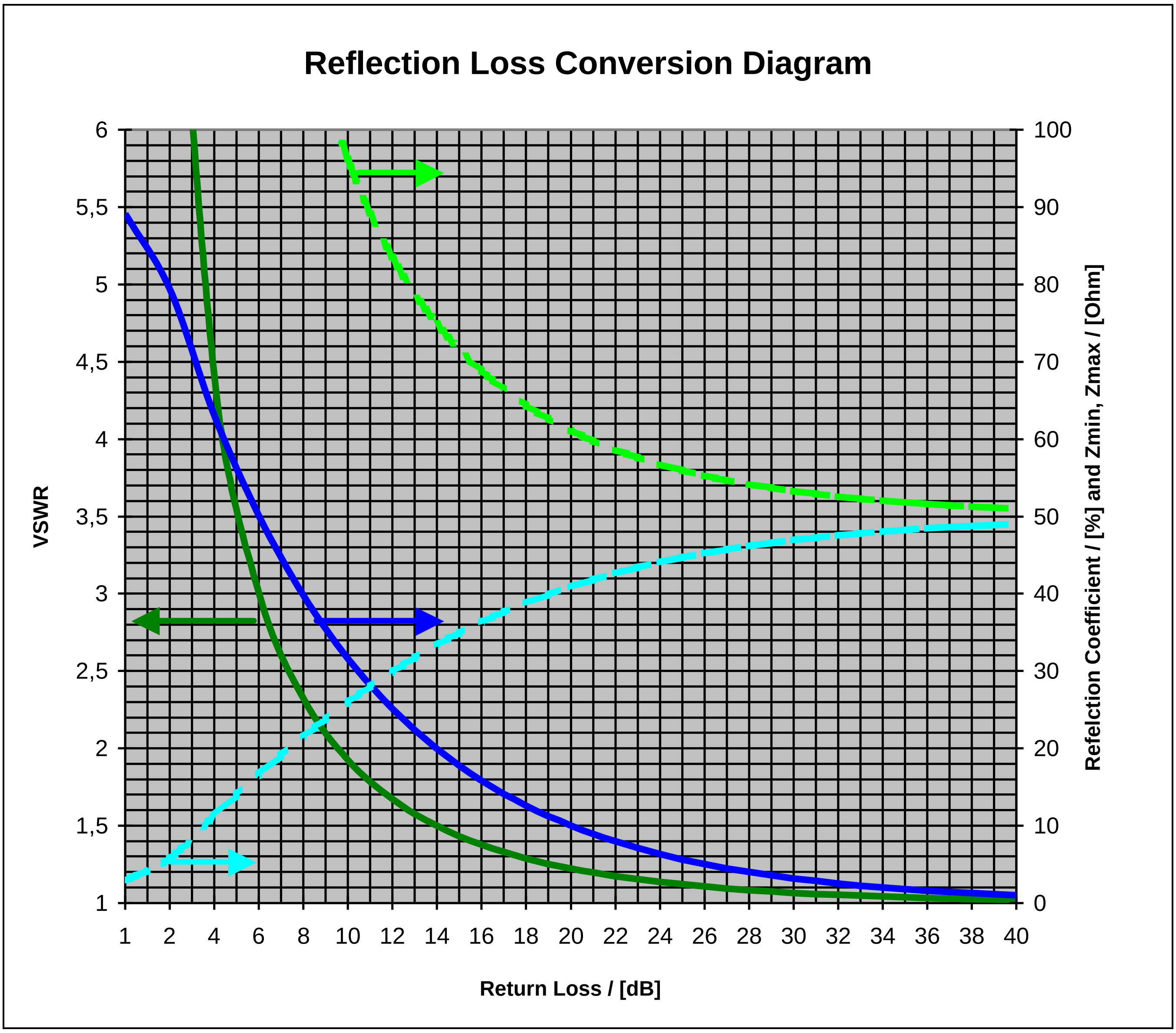
We can then examine *r*: $|r| = \left|\frac{96.25-50}{96.25+50}\right| = \left|\frac{25.96-50}{25.96+50}\right| = 0.316$

The λ/4 transformer transforms the device impedance to:

Z_{IN1}=96.25Ω → $Z_{Ende} = \frac{Z_O{}^2}{Z_{IN}} = \frac{50\Omega^2}{96.25\Omega} = 25.97\Omega$ and for Z_{IN2}=25.97Ω → 96.25Ω

Results: At 2GHz, the BGA2711 offers an input return loss of 10dB or VSWR=1.92. This reflection can be caused by a 96.25Ω or a 25.97Ω impedance. Of course there are infinite results possible if one takes into account all combinations of L and C values. Measuring this impedance at 2GHz with the use of a non-50Ω cable will cause extremely large errors in λ/4 distance, because the Zin_1 = 96.25Ω appears as 25.97Ω and the second solution Zin_2=25.97Ω appears as 96.25Ω!

As illustrated in the above example, the VSWR (or return loss) quickly associates the quality of device's input matching without any calculations, but does not tell about its real (vector) performance (missing of phase information). Detailed mathematical network analysis of RF amplifiers depends on the device's input impedance versus output load (S12 issue). The output device impedance is dependent on source's impedance driving the amplifier (S21 issue). Due to this interdependence, the use of s-parameters in linear small signal networks offers reliable and accurate results. This S-parameter theory will be presented in the next chapters.



Example: Select your interesting Return-Loss (10dB). Crossing the dark green trace you can find the VSWR (≈1,9) and crossing the dark blue trace you can find the Reflection Coefficient (r≈0,32). There are two (100% resistive) mismatches found either crossing the dashed light green traces (Z_{max}≈96Ω) or crossing the dashed light blue trace (Z_{min}≈26Ω). For further details, please referee to the former algebraic solved application example.

1.1.4 Difference between ideal and practical passive devices

Practical devices have so-called parasitic elements at very high frequencies.

Resistor Has an inductive parasitic action and acts like a low pass filtering function.
Inductor Has a capacitive and resistive parasitic, causing it to act like a damped parallel resonant tank circuit with a certain self resonance.
Capacitor Has an inductive and resistive parasitic, causing it to act like a damped tank circuit with ***S***eries ***R***esonance ***F***requency (***SRF***).

The inductor's and the capacitor's parasitic reactance causes self-resonances.

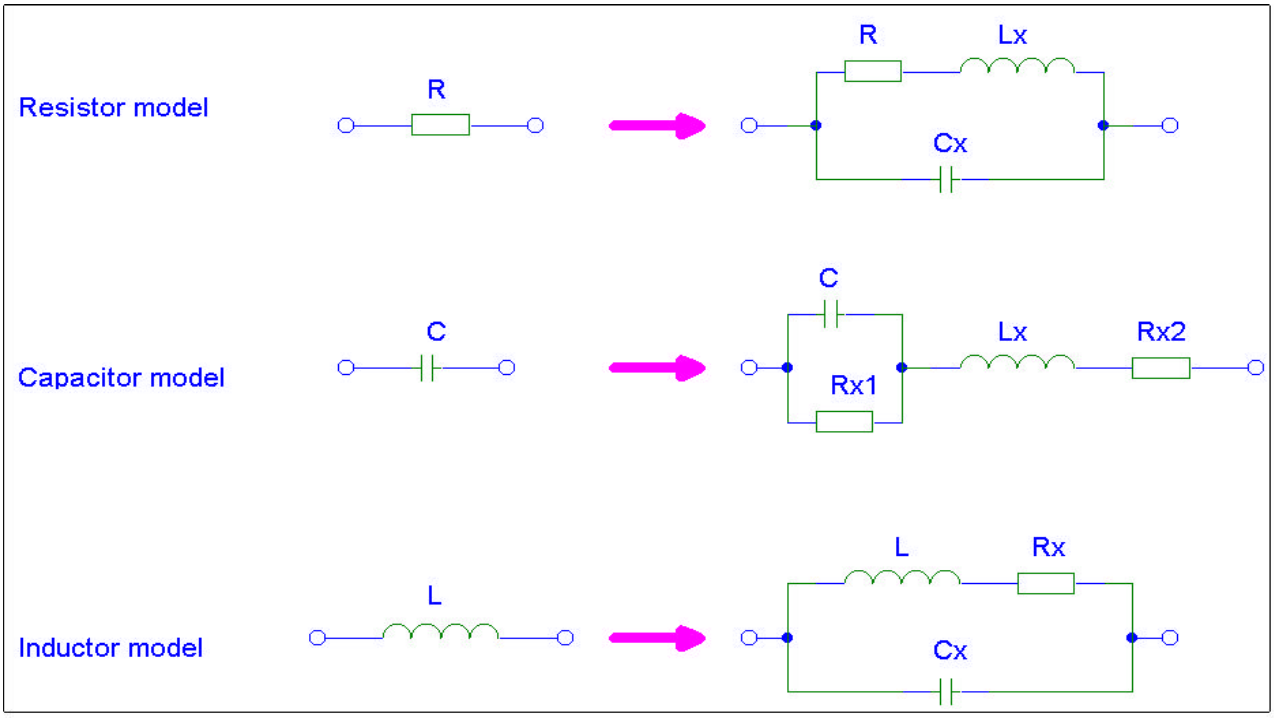


Figure 7: Equivalent models of passive lumped elements

The use of a passive component above its SRF is possible, but must be critically evaluated. A capacitor above its SRF appears as an inductor with DC blocking capabilities.

1.1.5 The Smith Chart

As indicated in an example in the former chapter, the impedances of semiconductors are a combination of resistive and reactive parts caused by phase delays and parasitic. RF is best analyzed in the frequency domain under the use of vector algebraic expressions:

Object	→	***into***	→	***Frequency domain***
Resistor	→	R	→	$R = \lvert R\rvert \cdot e^{+j0^\circ}$
Inductor	→	L	→	$X_L = +jw \cdot L = w \cdot L \cdot e^{+j90^\circ}$
Capacitor	→	C	→	$X_C = -j\frac{1}{w \cdot C} = \frac{1}{w \cdot C} \cdot e^{-j90^\circ}$
Frequency	→	f	→	$w = 2p \cdot f$
Complex designator	→	j	→	$+j = \sqrt{-1} = \frac{1}{-j} = e^{+j90^\circ}$

Some useful basic vector algebra in RF analysis:

Complex impedance:

$$\boxed{Z = \mathrm{Re}\{Z\} + j\,\mathrm{Im}\{Z\} = \lvert Z\rvert \cdot e^{jj} = \lvert Z\rvert \cdot (\cos j - j \sin j)}$$

$$\mathrm{Im}\{Z\} = \lvert Z\rvert \sin j \;;\; \mathrm{Re}\{Z\} = \lvert Z\rvert \cos j \;;$$

$\tan = \frac{\sin j}{\cos j}$ → $\tan j = \frac{\mathrm{Im}\{Z\}}{\mathrm{Re}\{Z\}}$; with $j = w \cdot t$

Use of angle → ***Polar*** notation
Use of sum → ***Cartesian (Rectangular)*** notation

The same rules are used for other issues,

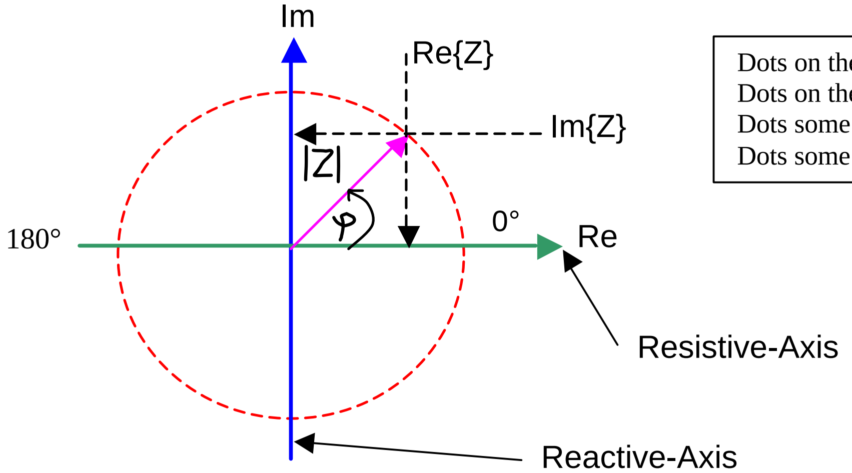
e.g., the ***complex reflection coefficient***:

$$\boxed{r = \lvert r\rvert \cdot e^{jj} = \frac{\lvert U_b\rvert \cdot e^{jj_b}}{\lvert U_f\rvert \cdot e^{jj_f}} = \left\lvert \frac{U_b}{U_f} \right\rvert \cdot e^{j(j_b - j_f)}}$$

Special cases:

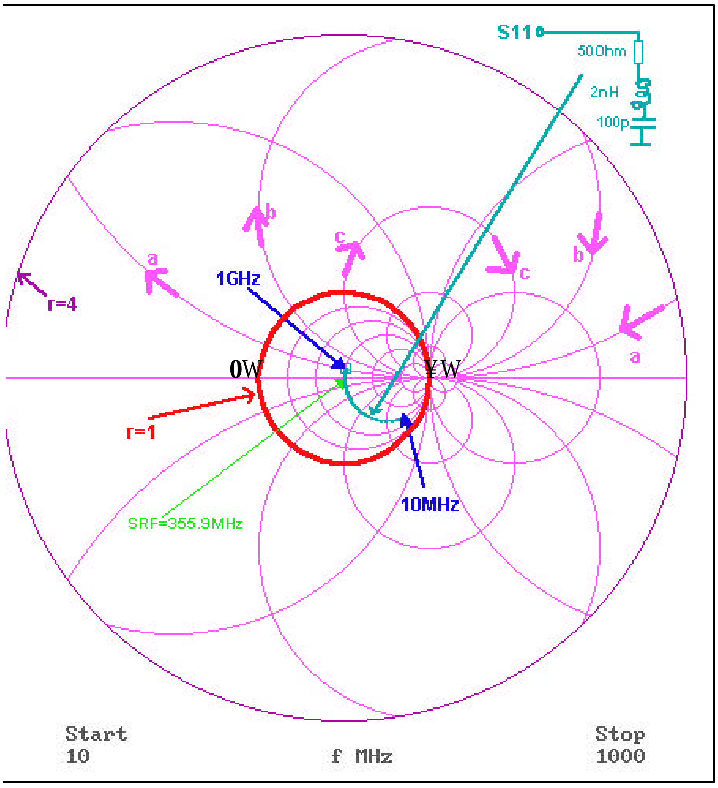
- Resistive mismatch: $j_{(R)} = 0^\circ$ reflection coefficient: $j_{(r)} = 0^\circ$
- Inductive mismatch: $j_{(L)} = +90^\circ$ reflection coefficient: $j_{(r)} = +90^\circ$
- Capacity mismatch: $j_{(C)} = -90^\circ$ reflection coefficient: $j_{(r)} = -90^\circ$

The ***Gaussian number area*** (***Polar Diagram***) allows charting rectangular two-dimensional vectors:



Dots on the Re-Line are 100% resistive
Dots on the Im-Line are 100% reactive
Dots some their above the Re-Line are inductive + resistive
Dots some their below the Re-Line are capacity + resistive

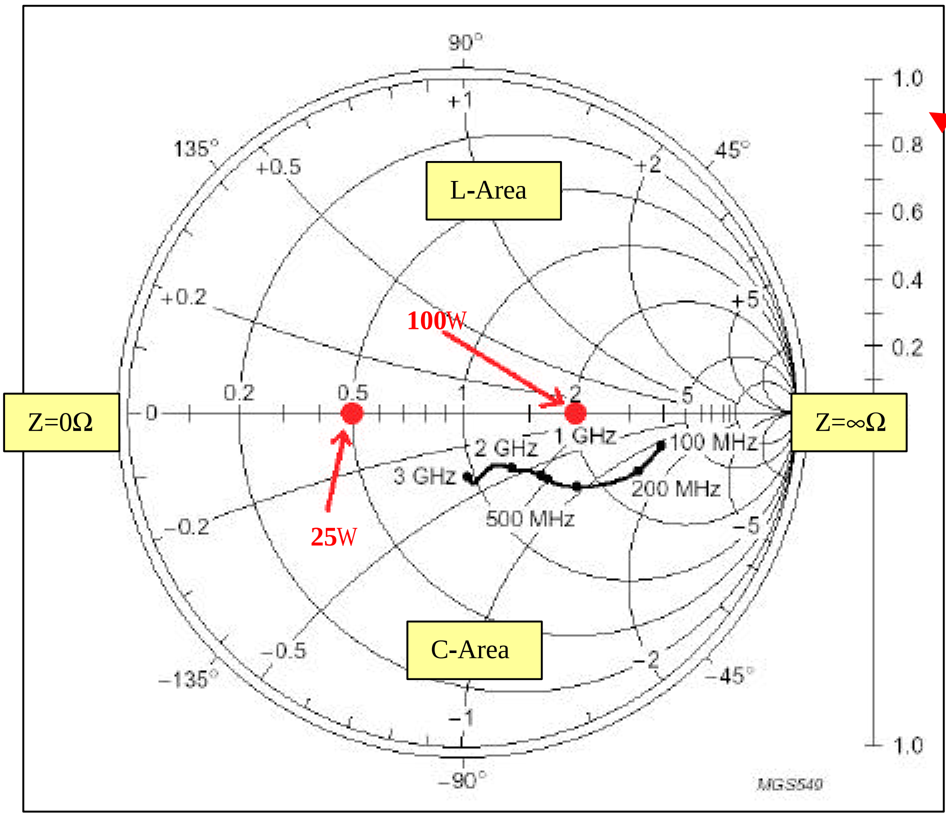
In applications RF designers try to remain close to a 50Ω resistive impedance. The polar diagram's origin is 0Ω. In RF circuit's, relative large impedances can occur but we try to remain close to 50Ω by special network design for maximum power transfer. Practically, very low and very high impedances don't need to be known accurately. The Polar diagram can't show simultaneous large impedances and the 50Ω region with acceptable accuracy, because of limited paper size.



Using this fact Mr. Phillip Smith, an engineer at Bell Laboratories developed in the 1930s the so-called ***Smith Chart***. The chart's origin is at 50Ω. Left and right resistive values along the real axis end in 0Ω and at ∞Ω. The imaginary reactive (imaginary axis, or Im-Axis) end ir 100% reactive (L or C). Close to the 50Ω origin high resolution is offered. Far away of the chart's centre does the resolution dope down. From the centre of the chart, the resolution / error increases. The standard Smith Char only displays ***positive resistances*** and has a unit radius (r=1). ***Negative resistances*** generated by ***instability*** (eg. ***oscillation***) lay outside the unit circle. In this non-linear scaled diagram, the infinite dot of the Re-Axis is "theoretically" bend to the zero point of the Smith Chart. Mathematically it can be shown that this will form the Smith Chart's unit circle (r=1). All dot's laying on this circle represent a reflection coefficient magnitude of 1 (100% mismatch). Any positive L/C combination with a resistor will be mathematically represented by it's polar notation reflection coefficient inside the Smith Chart's unity circle. Because the Smith Chart is a transformed linear scaled polar diagram we can use 100% of the polai diagram rules. The Cartesian-diagram rules are changed because of the non-linear scaling.

Special cases:

- Dots above the horizontal axis represents impedance with inductive part (0° < φ < 180°)
- Dots below the horizontal axis represents impedance with capacitive part (180° < φ < 360°)
- Dots laying on the horizontal axis (ordinate) are 100% resistive (φ = 0°)
- Dots laying on the vertical axis (abscissa) are 100% reactive (φ = 90°)



Scaling rule for determine the Magnitude (vector distance) of the reflection coefficient

Figure 8: BGA2003 output Smith Chart (S_{22})

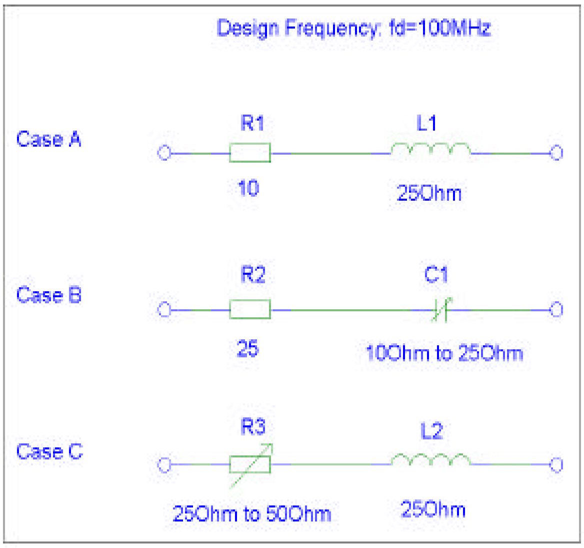
Illustrated are the special cases for ZERO and infinitely large impedance. The upper half circle is the inductive region. The lower half of the circle is the capacitive region. The origin is the 50Ω system reference (Z_O). To be more flexible, numbers printed in the chart are normalized to Z_O.

Normalizing impedance procedure: $Z_{norm} = \frac{Z_x}{Z_o}$ Z_O = System reference impedance (e.g., 50Ω, 75Ω)

Example: Plot a 100Ω & 50Ω resistor into the upper ***BGA2003***'s output Smith chart.
Calculation: Z_{norm1}=100Ω/50Ω=2; Z_{norm2}=25Ω/50Ω=0.5
Result: The 100Ω resistor appears as a dot on the horizontal axis at the location 2.
The 25Ω resistor appears as a dot on the horizontal axis at the location 0.5

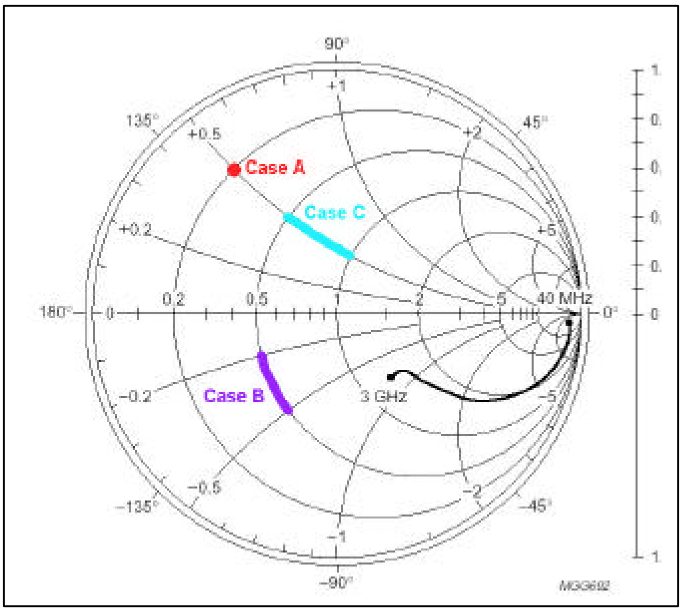
Example1: In the following three circuits, capacitors and inductors are specified by the amount of reactance @ 100MHz design frequency. Determine the value of the parts. Plot their impedance in to the ***BFG425W***'s output (S22) Smith Chart.

Circuit:



➔

Result:



Calculation:

Case A (constant resistance)

From the circuit ➔ $Z_A = 10\Omega + j25\Omega\,;\ L_1 = \dfrac{25\Omega}{2p \cdot 100MHz} = 39.8nH$

$Z_{(A)norm} = Z_A/50\Omega = 0.2 + j0.5$ ➔ Drawing into Smith Chart

Basics:

$$C = \frac{1}{w \cdot X_C}$$

$$L = \frac{X_L}{w}$$

$$w = 2p \cdot f$$

Case B (constant resistance and variable reactance - variable capacitor)

From the circuit ➔ $Z_B = 10\Omega + j(10 \text{ to } 25)\mathrm{O}$

$$C_B = \frac{1}{2p \cdot 100MHz \cdot (10 \text{ to } 25)\Omega} = 63.7\text{pF to } 159.2\text{pF}$$

$Z_{(B)norm} = Z_B/50\Omega = 0.5 - j(0.2 \text{ to } 0.5)$ ➔ Drawing into Smith Chart

Case C (constant resistance and variable reactance - variable inductor)

From the circuit ➔ $Z_C = (25\mathrm{O} \text{ to } 50\mathrm{O}) + j25\Omega\,;$

$$L_C = \frac{(25 \text{ to } 50)\Omega}{2p \cdot 100MHz} = 39.8\text{nH to } 79.6\text{nH}$$

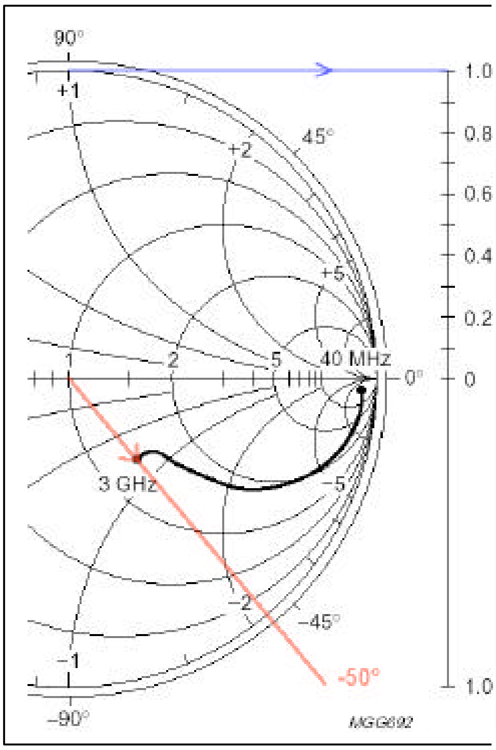
$Z_{(C)norm} = Z_C/50\Omega = (0.5 \text{ to } 1) + j0.5$ ➔ Drawing into Smith Chart

Example2: Determine ***BFG425W***'s outputs reflection coefficient (S22) at 3GHz from the data sheet. Determine the output return loss and output impedance. Compensate the reactive part of the impedance.

Calculation: Reading the data in the Smith Chart with improved resolution, is done by the use of the vector reflection coefficient in Polar notation.

Procedure: **1)** Mechanically measure the scalar length from the chart origin to the 3GHz (vector distance).
2) On the chart's right side is printed a ruler with the numbers of 0 to 1. Read from it the equivalent scaled scalar length |r| = 0.34
3) Measure the angle ∠(r) = φ = -50°. Write the reflection coefficient in vector polar notation

$r = 0.34e^{-j50°}$



Normalized impedance: $\frac{Z}{Z_O} = \frac{1+r}{1-r} = 1.513e^{-j30.5°}$

Because the transistor was characterized in a 50Ω bench test set-up ➔ Z_o = 50Ω

Impedance: $Z_{22} = 75.64\Omega e^{-j30.5°} = (65.2 - j38.4)\Omega$

$$C = \frac{1}{2p \cdot 3GHz \cdot 38.4\Omega} = 1.38\,pF$$

The output of BFG425W has an equivalent circuit of 65.2Ω with 1.38pF series capacitance. Output return loss, not compensated: RL_{OUT}= -20log(|r|)=9.36dB resulting in $VSWR_{OUT}$=2

For compensation of the reactive part of the impedance, we take the ***conjugate complex*** of the reactance:

Xcon=-Im{Z} = -{-j38.4Ω} = +j38.4Ω resulting in $L = \frac{38.4\Omega}{2p \cdot 3GHz} = 2nH$

A 2nH series inductor will compensate the capacitive reactance.

The new input reflection coefficient is calculated to: $r = \frac{65.2\Omega - 50\Omega}{65.2\Omega + 50\Omega} = 0.132$

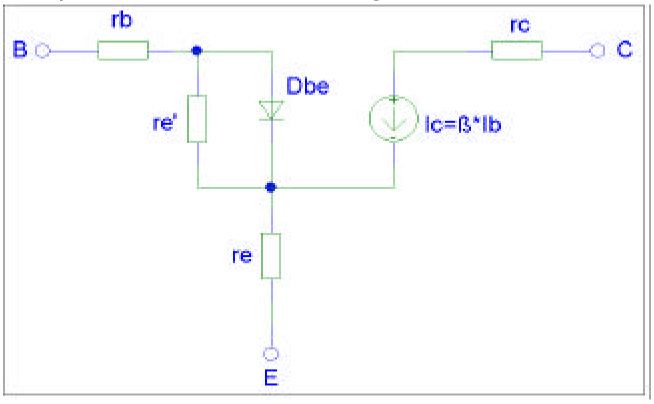
Output return loss, compensated: RL_{OUT}= -20log(0.132)=17.6dB resulting in $VSWR_{OUT}$=1.3

Please note: In practical situations the output impedance is a function of the input circuit. The input and output matching circuits are defined by the ***stability*** requirements, the need gain and noise-matching. Investigation is done by using network analysis based on ***S-Parameters***.

1.2 Small signal RF amplifier parameters

1.2.1 Transistor parameters, DC to microwave

At low DC currents and voltages, one can assume a transistor acts like a voltage-controlled current source with diode clamping action in the *base-emitter* input circuit. In this ***model***, the transistor is specified by its large signal DC-parameters, i.e., DC-current gain (B, ß, h_{fe}), maximum power dissipation, breakdown voltages and so forth.



$$I_C = I_{CO} \cdot e^{\frac{U_{BE}}{V_T}} \qquad r_e' = \frac{V_T}{I_E}$$

Thermal Voltage: $V_T = kT/q \approx 26mV@25°C$

I_{CO}=Collector reverse saturation current

Low frequency voltage gain: $V_u \approx \frac{R_C}{r_e'}$

Current gain $ß = \frac{I_C}{I_B}$

Increasing the frequency to the audio frequency range, the transistor's parameters get frequency-dependent phase shift and parasitic capacitance effects. For characterization of these effects, small signal ***h-parameters*** are used. These hybrid parameters are determined by measuring voltage and current at one terminal and by the use of open or short (standards) at the other port.

The ***h-parameter*** matrix is shown below.

h-Parameter Matrix: $\begin{pmatrix} u_1 \\ i_2 \end{pmatrix} = \begin{pmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{pmatrix} * \begin{pmatrix} i_1 \\ u_2 \end{pmatrix}$

Increasing the frequency to the HF and VHF ranges, open ports become inaccurate due to electrically stray field radiation. This results in unacceptable errors. Due to this phenomenon ***y-parameters*** were developed. They again measure voltage and current, but use of only a "short" standard. This "short" approach yields more accurate results in this frequency region. The ***y-parameter*** matrix is shown below.

y-Parameter Matrix: $\begin{pmatrix} i_1 \\ i_2 \end{pmatrix} = \begin{pmatrix} y_{11} & y_{12} \\ y_{21} & y_{22} \end{pmatrix} * \begin{pmatrix} u_1 \\ u_2 \end{pmatrix}$

Further increasing the frequency, the parasitic inductance of a "short" causes problem due to mechanical depending parasitic. Additionally, measuring voltage, current and it's phase is quite tricky. The scattering parameters, or ***S-parameters,*** were developed based on the measurement of the forward and backward traveling waves to determine the reflection coefficients on a transistor's terminals (or ports). The ***S-parameter*** matrix is shown below.

S-Parameter Matrix: $\begin{pmatrix} b_1 \\ b_2 \end{pmatrix} = \begin{pmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{pmatrix} * \begin{pmatrix} a_1 \\ a_2 \end{pmatrix}$

1.2.2 Definition of the S-Parameters

Every amplifier has an input port and an output port (a 2-port network). Typically the input port is labeled Port-1 and the output is labeled Port-2.

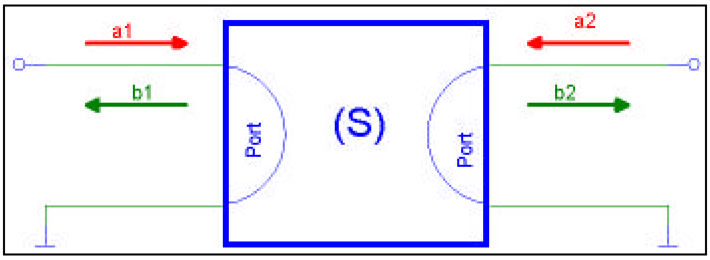


Figure 10: Two-port Network's (a) and (b) waves

Matrix: $\begin{pmatrix} b_1 \\ b_2 \end{pmatrix} = \begin{pmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{pmatrix} * \begin{pmatrix} a_1 \\ a_2 \end{pmatrix}$

Equation: $b_1 = S_{11} \cdot a_1 + S_{12} \cdot a_2$

$b_2 = S_{21} \cdot a_1 + S_{22} \cdot a_2$

The forward-traveling waves **(a)** are traveling into the DUT's (input or output) ports.
The backward-traveling waves **(b)** are reflected back from the DUT's ports
The expression "port Z_O terminate" means the use of a 50Ω-standard. This is not a conjugate complex power match! In the previous chapter the reflection coefficient was defined as:

Reflection coefficient: $r = \frac{back\ running\ wave}{forward\ running\ wave}$

Calculating the ***input reflection factor*** on port 1: $S_{11} = \frac{b_1}{a_1}\Big|_{a_2=0}$ with the output terminated in Z_O.

That means the source injects a forward-traveling wave **(a1)** into Port-1. No forward-traveling power **(a2)** injected into Port-2. The same procedure can be done at Port-2 with the

Output reflection factor: $S_{22} = \frac{b_2}{a_2}\Big|_{a_1=0}$ with the input terminated in Z_O.

Gain is defined by: $gain = \frac{output\ wave}{input\ wave}$

The ***forward-traveling wave gain*** is calculated by the wave **(b2)** traveling out off Port-2 divided by the wave **(a1)** injected into Port-1.

$$S_{21} = \frac{b_2}{a_1}\Big|_{a_2=0}$$

The ***backward traveling wave gain*** is calculated by the wave **(b1)** traveling out off Port-1 divided by the wave **(a2)** injected into Port-2. $S_{12} = \frac{b_1}{a_2}\Big|_{a_1=0}$

The normalized waves (a) and (b) are defined as:

$$a_1 = \frac{1}{2\sqrt{Z_O}}(V_1 + Z_O \cdot i_1) \quad = \quad \text{signal into Port-1}$$

$$a_2 = \frac{1}{2\sqrt{Z_O}}(V_2 + Z_O \cdot i_2) \quad = \quad \text{signal into Port-2}$$

$$b_1 = \frac{1}{2\sqrt{Z_O}}(V_1 + Z_O \cdot i_1) \quad = \quad \text{signal out of Port-1}$$

$$b_2 = \frac{1}{2\sqrt{Z_O}}(V_1 + Z_O \cdot i_2) \quad = \quad \text{signal out of Port-2}$$

The normalized waves have units of $\sqrt{Watt}$ and are referenced to the system impedance Z_O. It is shown by the following mathematical analyses:
The relationship between U, P an Z_O can be written as:

$$\frac{u}{\sqrt{Z_O}} = \sqrt{P} = i \cdot \sqrt{Z_O} \qquad \text{Substituting: } \frac{Z_0}{\sqrt{Z_O}} = \sqrt{Z_O}$$

$$a_1 = \frac{V_1}{2\sqrt{Z_O}} + \frac{Z_O \cdot i_1}{2\sqrt{Z_O}} = \frac{\sqrt{P_1}}{2} + \frac{Z_O \cdot i_1}{2\sqrt{Z_O}}$$

$$a_1 = \frac{\sqrt{P_1}}{2} + \frac{\sqrt{Z_O} \cdot i_1}{2} = \frac{\sqrt{P_1}}{2} + \frac{\sqrt{P_1}}{2} \rightarrow a_1 = \sqrt{P_1} \quad (\rightarrow \text{Unit} = \sqrt{Watt} = \frac{Volt}{\sqrt{Ohm}})$$

Forward transmission:
$FT = 20\log(S_{21})dB$

Isolation:
$S12(dB) = -20\log(S_{12})dB$

Input Return Loss:
$RL_{in} = -20\log(S_{11})dB$

Output Return Loss:
$RL_{OUT} = -20\log(S_{22})dB$

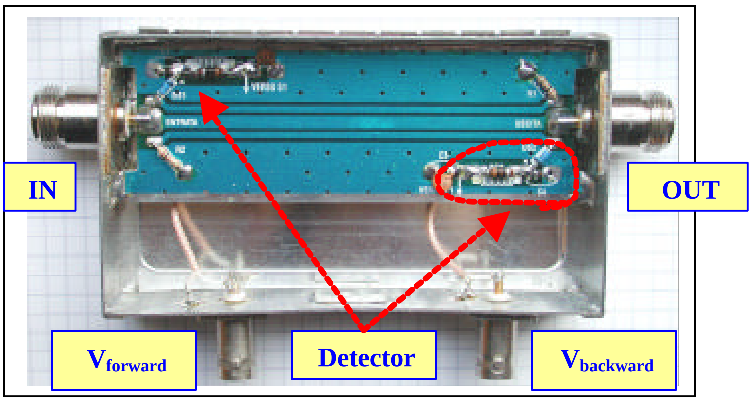
Insertion Loss:
$IL = -20\log(S_{21})dB$

Rem:

$$\frac{Z_O}{\sqrt{Z_O}} = \frac{Z_O \cdot \sqrt{Z_O}}{\sqrt{Z_O} \cdot \sqrt{Z_O}} = \frac{Z_O \cdot \sqrt{Z_O}}{Z_O} = \sqrt{Z_O}$$

$$P = U \cdot I = \frac{U^2}{R} \rightarrow \sqrt{P} = \frac{U}{\sqrt{R}} = I \cdot \sqrt{R}$$

Because $a_1 = \frac{V_{forward}}{\sqrt{Z_O}}$, the normalized waves can be determined the measuring the voltage of a forward-traveling wave referenced to the system impedance constant $\sqrt{Z_O}$. Directional couplers or VSWR bridges can divide the standing waves into the forward- and backward-traveling voltage wave. (Diode) Detectors convert these waves to the $V_{forward}$ and $V_{backward}$ DC voltage. After an easy processing of both DC voltages, the VSWR can be read.



50Ω VHF-SWR-Meter built from a kit (Nuova Elettronica). consists of three strip-lines. The middle line passes the main signal from the input to the output. The upper and lower stri lines select a part of the forward and backward traveling wav by special electrical and magnetic cross-coupling. Diode detectors at each coupled strip-line-end rectify the power to a DC voltage, which is passed to an external analog circuit for processing and monitoring of the VSWR. Applications: Pow antenna match control, PA output power detector, vector voltmeter, vector network analysis, AGC, etc. These kinds o circuit's kits are published in amateur radio literature and in several RF magazines.

1.2.2.1 2-Port Network definition

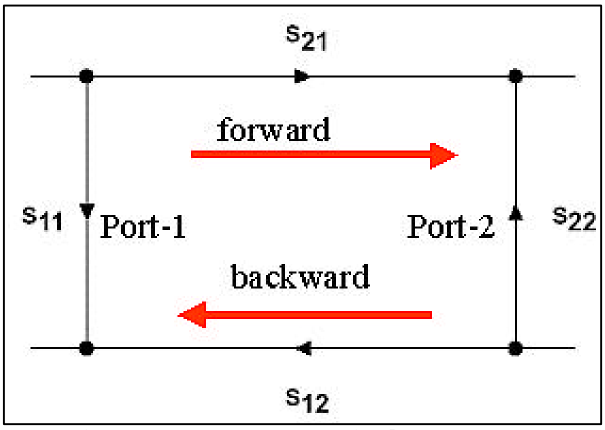


Figure 11: S-Parameters in the Two-port Network

Input return loss

$$S_{11} = \sqrt{\frac{\text{Power reflected from input port}}{\text{Power available from generator at input port}}}$$

Output return loss

$$S_{22} = \sqrt{\frac{\text{Power reflected from output port}}{\text{Power available from generator at output port}}}$$

Forward transmission loss (insertion loss)

$$S_{21} = \sqrt{\text{Transducer power gain}}$$

Reverse transmission loss (isolation)

$$S_{12} = \sqrt{\text{Reverse transducer power gain}}$$

Philips' data sheet parameter ***Insertion power gain*** $|S_{21}|^2$: $10dB \cdot \log|S_{21}|^2 = 20dB \cdot \log|S_{21}|$

Example: Calculate the insertion power gain for the ***BGA2003*** at 100MHz, 450MHz, 1800MHz, and 2400MHz for the bias set-up $V_{VS\text{-}OUT}$=2.5V, $I_{VS\text{-}OUT}$=10mA.

Calculation: Download the S-Parameter data file [2_510A3.S2P] from the Philips' website page for the Silicon MMIC amplifier BGA2003.

This is a section of the file:

\# MHz S MA R 50

! Freq	S11		S21		S12		S22	:
100	0.58765	-9.43	21.85015	163.96	0.00555	83.961	0.9525	-7.204
400	0.43912	-28.73	16.09626	130.48	0.019843	79.704	0.80026	-22.43
500	0.39966	-32.38	14.27094	123.44	0.023928	79.598	0.75616	-25.24
1800	0.21647	-47.97	4.96451	85.877	0.07832	82.488	0.52249	-46.31
2400	0.18255	-69.08	3.89514	76.801	0.11188	80.224	0.48091	-64

Results:

100MHz ➔ 20·log(21.85015) = 26.8 dB

450MHz ➔ $20dB\log\left|\frac{16.09626e^{130.48°}+14.27094e^{123.44°}}{2}\right| = 23.6dB$

1800MHz ➔ 20·log(4.96451) = 13.9 dB

2400MHz ➔ 20·log(3.89514) = 11.8 dB

1.2.2.2 3-Port Network definition

Typical vehicles for 3-port s-parameters are: Directional couplers, power splitters, combiners, and phase splitters.

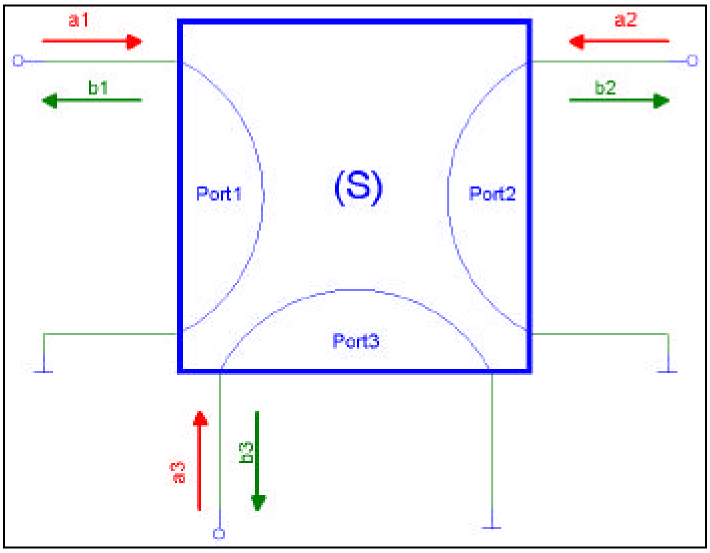


Figure 12: Three-port Network's (a) and (b) waves

3-Port s-parameter definition:

- Port reflection coefficient / return loss:

Port 1 ➔ $S_{11} = \frac{b_1}{a_1}|_{(a_2=0;\, a_3=0)}$

Port 2 ➔ $S_{22} = \frac{b_2}{a_2}|_{(a_1=0;\, a_3=0)}$

Port 3 ➔ $S_{33} = \frac{b_3}{a_3}|_{(a_1=0;\, a_2=0)}$

- Transmission gain:

Port 1=>2 ➔ $S_{21} = \frac{b_2}{a_1}|_{(a_3=0)}$

Port 1=>3 ➔ $S_{31} = \frac{b_3}{a_1}|_{(a_2=0)}$

Port 2=>3 ➔ $S_{32} = \frac{b_3}{a_2}|_{(a_1=0)}$

Port 2=>1 ➔ $S_{12} = \frac{b_1}{a_2}|_{(a_3=0)}$

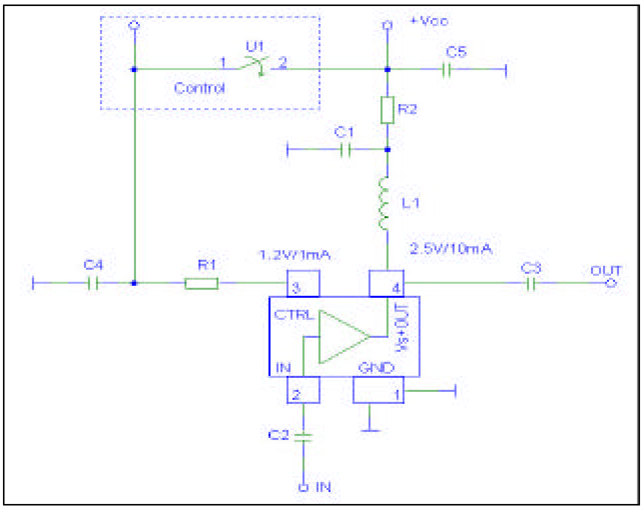
Port 3=>1 ➔ $S_{31} = \frac{b_1}{a_3}|_{(a_2=0)}$

Port 3=>2 ➔ $S_{23} = \frac{b_3}{a_2}|_{(a_1=0)}$

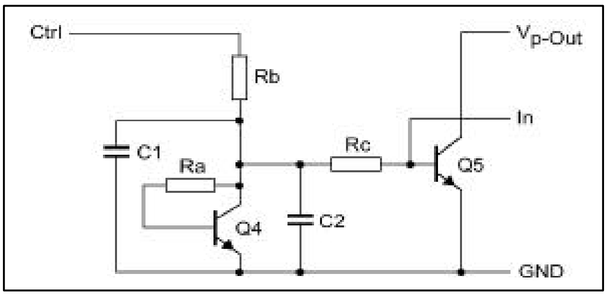
1.3 RF Amplifier design Fundamentals

1.3.1 DC bias point adjustment at MMICs

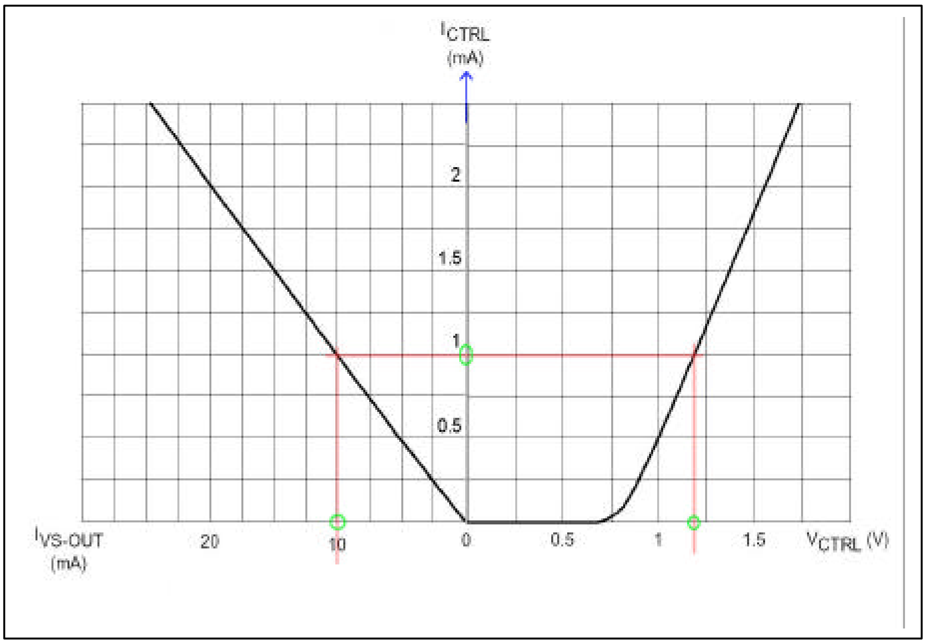
As shown in the former chapter S-Parameters are depending on the bias point and the frequency. Due to it S-Parameter files do include the dc bias-setting data. It's recommended of using this setup because the S-Parameter will not be valid for a different bias point. An example of principle dc bias circuit design is illustrate on ***BGU2003*** for Vs=2.5V; Is=10mA. The supply voltage is choose to be V_{CC}=3V.



Principal LNA DC bias setup



BGA2003 equivalent circuit: Q_5 is the main RF transistor. Q_4 forms a current mirror with Q_5. The input current of this current mirror is determined by the current into pin Ctrl. R_b limits the current when a control voltage is applied directly to the Ctrl input. R_C, C_1, and C_2 decouple the bias circuit from the RF input signal. Because Q_4 and Q_5 are located on the same die, Q_5's bias point is very temperature stable.



BGU2003: Device I/O DC bias setup via PIN3 (CTRL)

From the BGA2003 datasheet were combined Figure 4 and Figure 5 into the adjacent picture for improved illustration of the MMICs I/O dc relationship.

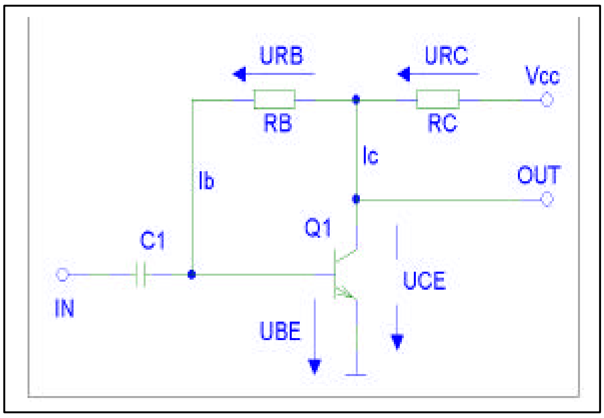
The red line shows the graphically construction starting with the need of $I_{VS\text{-}OUT}$=10mA. Automatically crossing the ordinate I_{CTRL}=1mA finishing into the abscissa at V_{CTRL}=1.2V

$$R_2 = \frac{V_{cc} - V_S}{I_{VS-OUT}} = \frac{3V - 2.5V}{10mA} = 50\Omega$$

$$R_1 = \frac{V_{cc} - V_{CTRL}}{I_{VS-OUT}} = \frac{3V - 1.2V}{1mA} = 1.8k\Omega$$

1.3.2 DC bias point adjustment at Transistors

As a contrast to the easy bias setup at MMICs, there is shown the design of a setup used at eg. audio or IF amplifiers.



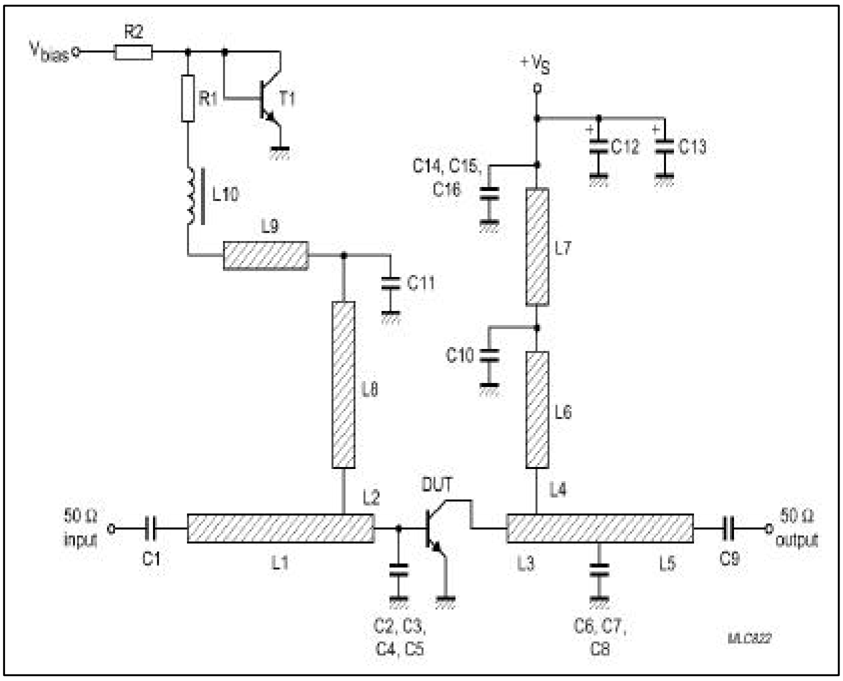
$$h_{FE} = b = B = \frac{I_C}{I_b} \qquad R_C = \frac{V_{CC} - U_{CE}}{I_b + I_C} = \frac{V_{CC} - U_{CE}}{I_C\left(\frac{h_{FE}+1}{h_{FE}}\right)}$$

$$\boxed{R_C = \frac{V_{CC} - U_{CE}}{I_C(h_{FE}+1)} \cdot h_{FE}};\ V_{CC} - I_C \cdot R_C = V_{CE} = I_b \cdot R_B + U_{BE}$$

$$\boxed{R_B = h_{FE} \cdot \left(\frac{V_{CC} - U_{BE}}{I_C} - R_C\right) = h_{FE} \cdot \frac{V_{CE} - V_{BE}}{I_C}}$$

DC bias setup with stabilization via Voltage feedback

The advantage is a very high resistive resistor R_B. It's lowering of the input impedance at terminal [IN] can be negated. Due to it a possible picked up IF-Band filter is less loaded. Due to missing of the emitter feedback resistor high gain is achieve from Q1. This is need for narrow bandwidth high gain IF amplifiers. The disadvantage is a very low stability of the operating point caused by the Si BE-Diodes' relative linear negative temperature coefficient of ca.$V_{BE} \approx$-2.5mV/K into amplified $I_C = \frac{V_C - V_{BE}}{R_B} \cdot h_{FE}$ This can be lowered by adding an additionally resistor between ground and the emitter.



An emitter resistor has the disadvantage of gain loss or the need of a bypassing capacitor. Additionally the transistor is loosing quality of gnd performances (instability) and an emitter heat sinking into the gnd plane. At medium output power, the bias setup must be stabilized due to the increased junction temperature causing dc drifting. Without stabilization the transistor will burn out or distortion can rise up. A possible solution is illustrated in the adjacent picture (***BFG10***). Comparable to the BGA2003, a current mirror is designed together with the dc transistor T1. T1 works like a diode with a V_{CE} (V_{BE}) drift close to the RF transistor (DUT) in the case of close thermal coupling. With $\beta_1=\beta_{DUT}$ and $V_{BE-1} \approx V_{BE-DUT}$ we can do a very simplified algebraic analysis:

$$V_T \cdot \ell n\left(\frac{I_{C-1}}{I_{CO}}\right) \approx V_T \cdot \ell n\left(\frac{I_{C-DUT}}{I_{CO}}\right)$$

finalizing into a very temperature independent relation ship of $I_{C\text{-}DUT} \approx I_{C1} \approx (V_{bias}-V_{BE})/R_2$
For best current imaging the BE die structure areas should have similar dimensions.

1.3.3 Gain Definitions

The gain of an amplifier is specified on several ways depending on how the (theoretically) measurement is implemented and depending on stability conditions, way of matching for e.g. best power processing, max. gain, lowest noise figure or a certain stability performance. Often certain power gain's are calculated for the upper and lower possible extreme of validity range. Additionally calculating circles in the smith chart (power gain circles, stability circles) are possible for selecting a useful working range in the input or output. The used algebraic expressions can vary from one literature source to the other one. In reality the S12 cannot be neglecting, causing the output being a function of the need source and the input being a function of the need load. This makes matching complicated and is a part of the G_A and G_P design procedure.

Transducer power gain:

$$G_T = \frac{P_L}{P_{AVS}} = \frac{\text{power delivered to the load}}{\text{power available from the source}}$$

Including the effect of I/O matching and device gain. Don't take into account the losses in components.

Power gain or operating power gain:

$$G_P = \frac{P_L}{P_{IN}} = \frac{\text{power deliverd to the load}}{\text{power input to the network}}$$

Used in the case of non neglect able S_{12} G_P is independent of the source impedance.

Available power gain:

$$G_A = \frac{P_{AVN}}{P_{AVS}} = \frac{\text{power avaialble from the network}}{\text{power avaialble from the source}}$$

G_A is independent of the load impedance.

Maximum available gain:

$$MAG = G_{T,\max} = 10\log\left(\frac{|S_{21}|}{|S_{12}|}\cdot\left|K \pm \sqrt{K^2 - 1}\right|\right)$$

The MAG you could ever hope to get from a transistor is under simultaneous conjugated I/O match with a ***Rollett Stability*** factor of K>1. K is calculated from the S-parameters in several sub steps. At frequency of unconditional stability, MAG ($G_{T,max}$=$G_{P,max}$=$G_{A,max}$) is plotted in transistor data sheets.

Maximum stable gain:

$$MSG = \frac{|S_{21}|}{|S_{12}|}$$

MSG is a figure of merit for a potential unstable transistor and valid for K=1 (subset of MAG).
At frequency of potential instability, MSG is plotted in transistor data sheets.

Further examples of used definitions in the design of amplifiers:
- $G_{T,\ max}$ = Maximum transducer power gain under simultaneous conjugated match conditions
- $G_{T,min}$ = Minimum transducer power gain under simultaneous conjugated match conditions
- G_{TU} = Unilateral transducer power gain
- $G_{P,min}$ = Minimum operating power gain for potential unstable devices
- Unilateral figure of merit $\frac{G_T}{G_{TU}}$ determine the error caused by assuming S_{12}=0.

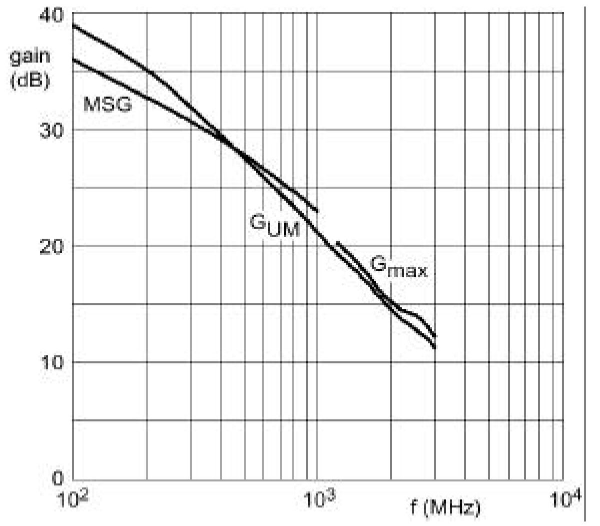
As an example is sown in the adjacent picture
the BGU2003 gain as a function of frequency

In the frequency range of 100MHz to 1GHz the MMIC is potential unstable. Above of ca. 1.2GHz the MMIC is unconditionally stable (within 3GHz range of measurement)

G_{UM} is the maximum unilateral transducer power gain assuming S_{12}=0 and a conjugated I/O match:
A S_{12}=0 (=unilateral figure of merit) specify an unilateral 2-port network resulting in K=infinite and $D_S=S_{11}*S_{22}$

$$G_{TU,\max} = \frac{|S_{21}|^2}{\left(1-|S_{11}|^2\right)\cdot\left(1-|S_{22}|^2\right)}$$

$$G_{UM} = 10\log\left(G_{TU,\max}\right)$$



($V_{S\text{-}OUT}$=2.5V; $I_{VS\text{-}OUT}$=10mA)

For further details please referee to books of e.g. Pozar, Gonzalez, Bowick and a lot of other guys.

1.3.4 Amplifier stability

All variables must be processed with complex data. The evaluated K-Factor is only valid for the frequency and bias setup for the selected S-parameter quartet [S_{11}, S_{12}, S_{21}, S_{22}]

Determinant: $D_S = S_{11}\cdot S_{22} - S_{12}\cdot S_{21}$

Rollett Stability Factor: $$K = \frac{1+|D_S|^2-|S_{11}|^2-|S_{22}|^2}{2\cdot|S_{21}\cdot S_{12}|}$$

In some literature sources the amount of D_S isn't take into account for the dividing into the following stability qualities.

K>1 & |Ds|<1
Unconditionally stable for any combination of source and load impedance

K<1
Potentially unstable and will most likely oscillate with certain combinations of source and load impedance. It does not mean that the transistor will not be useable for the application. It means the transistor is more tricky in use. A simultaneous conjugated match for the I/O isn't possible.

-1<K<0
used in oscillator designs

K>1 & |Ds|>1
This potentially unstable transistors with the need $SWR_{(IN)}=SWR_{(OUT)}=1$ are not manufactured and do have a gain of $G_{T,min}$.

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Appendix D: Application of the RF Switch BF1107/8 Mosfets

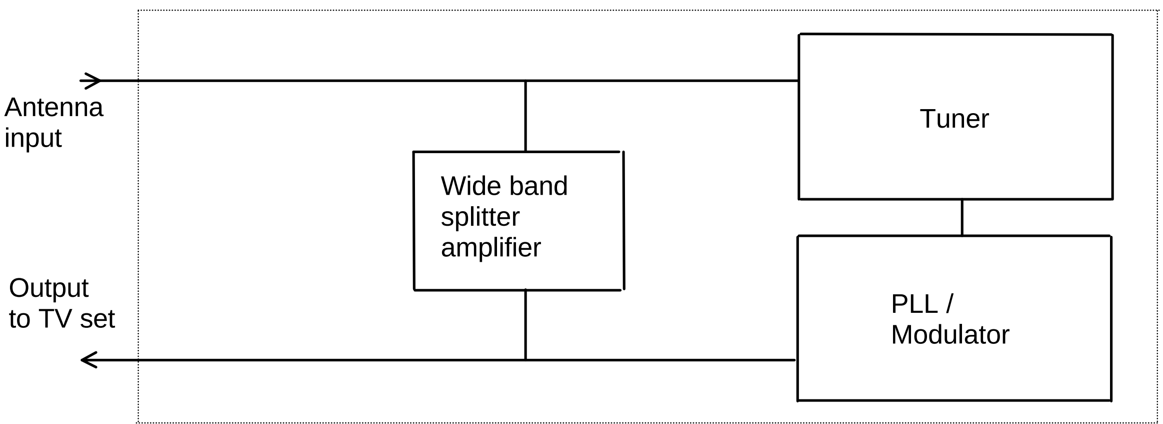
APPLICATION OF THE RF SWITCH BF1107

INTRODUCTION

If a (Mos)fet is used in its linear region, it can be used as a variable resistor. The resistance depends on the bias voltage between Gate and Source and the pinch - off voltage of the Mosfet.
If the bias voltage is lower than the pinch - off voltage the resistance of the Mosfet is infinite. If the bias voltage is much higher than the pinch - off voltage the resistance of the Mosfet is low.
Due to this a Mosfet can be used as a switch.
At low Gate - Source voltages the Mosfet is switched off and at high Gate - Source voltages the Mosfet is switched on.
If a Mosfet is used with relatively low capacitances the Mosfet can be used as an RF switch. With this Rf switch, RF signals can be switched off and on.
The BF1107 is a triode Mosfet intended for switching RF signals.
If the Drain - Source voltage is set to 0V, this Mosfet is biased in its linear region.
This Mosfet has a pinch - off voltage of approx. 3V.
Therefore this Mosfet is switched on if the Gate - Source voltage is 0V. Together with a Drain - Source voltage of 0V this means that the Mosfet is switched on if all bias voltages are 0V.
If the Gate - Source voltage is set to a value lower than 3V this Mosfet is switched off.

APPLICATION IN A VIDEO RECORDER

A block diagram of the principle circuit of the RF front end of a VCR is given in Fig.1 below.



If the VCR is not used ("stand-by") at least the wide band splitter amplifier must always be switched on to ensure reception of TV signals in the TV set.

Power consumption in stand-by can be reduced if the supply voltage of the VCR can be switched off, but special measures must be taken to ensure the reception of TV signals.
This can be done by connecting a switch between the input and output. (See Fig. 2 below). This is a so called “Passive Loop Through”.

To reduce power consumption the switch must be:
- on if the VCR is switched off and
- off of the VCR is switched on.

If for the switch a depletion type Mosfet is chosen then this Mosfet is switched on if all the supply voltages at the Mosfet are 0.
The Mosfet is switched off if the Gate - Source voltage has a negative value more negative than the pinch-off voltage of the Mosfet.

If the supply voltage of the VCR is switched on the Mosfet switch must be switched off. This can be done by connecting the Drain and the Source of the Mosfet to the supply voltage and connecting the Gate to ground.
The principle of this is given in Fig. 4 (next page).
If the supply voltage = 0, than the Drain-, Source- and Gate voltages of the Mosfet switch are 0. Than the antenna signal flows through the Mosfet switch to the TV set. If the supply voltage = 5V, then the Drain and Source voltages of the Mosfet switch are 5V. The capacitor C ensures that the Drain and the Source voltages are equal. The Gate voltage is 0 (Gate is grounded).
Then the antenna signal flows through the VCR as usual.

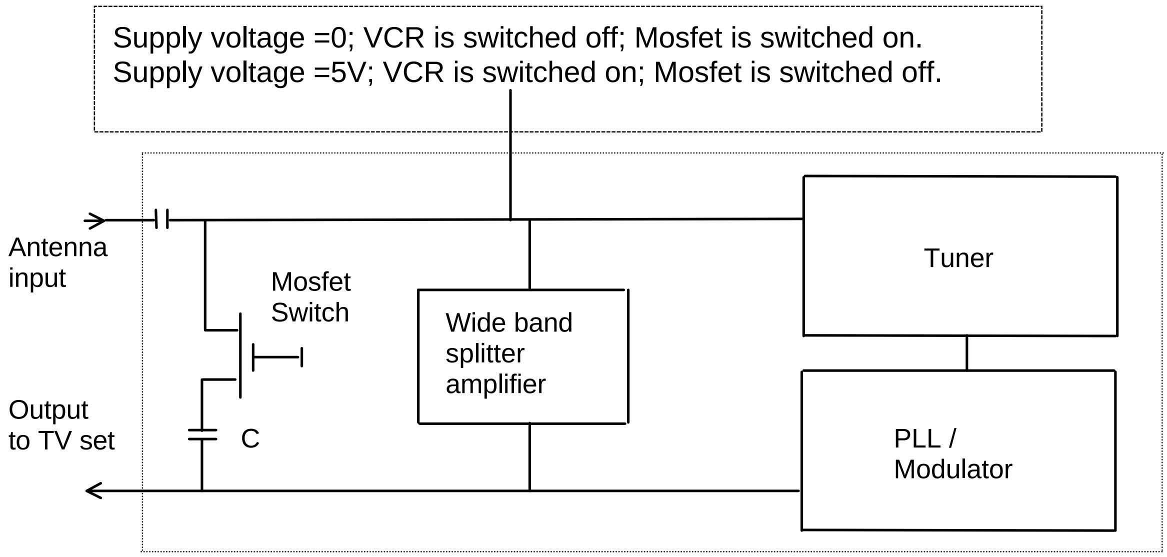


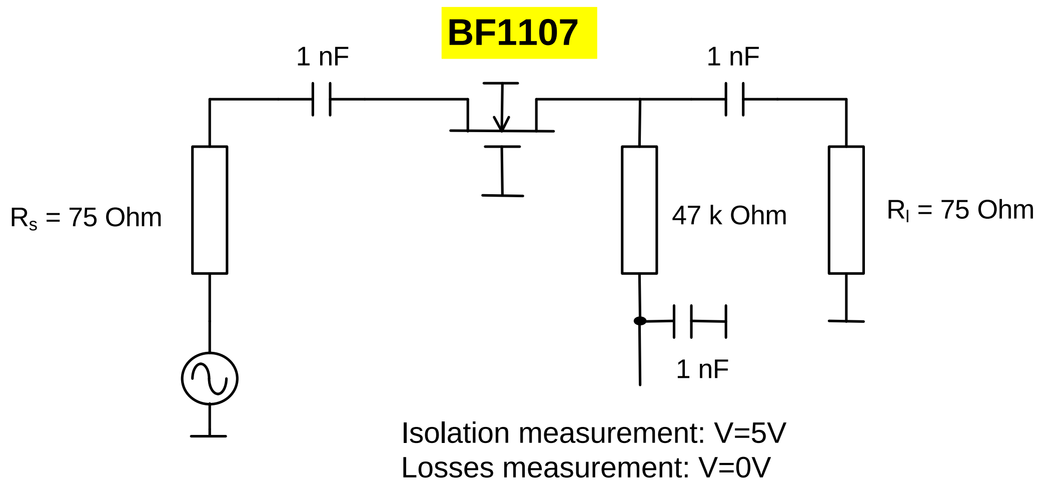
Fig. 4

For the Mosfet switch in this circuit a BF1107 can be applied. In the on state of the switch the losses must be low, because losses determine, for a large amount, the increase of the noise figure of the TV set. In the off state the

isolation must be high because the oscillator signal from the modulator must be kept very small at the antenna input.
The main advantage of applying the BF1107 as a switch for the passive loop through is that this Mosfet uses no current. Not in the on state, nor in the off state. Switching is done only with voltages.

PERFORMANCE OF THE BF1107

The performance of the RF switch was measured in a circuit as given in Fig. 5.



In this circuit we measured isolation and losses as a function of frequency.
The results of these measurements are given in Fig. 6.

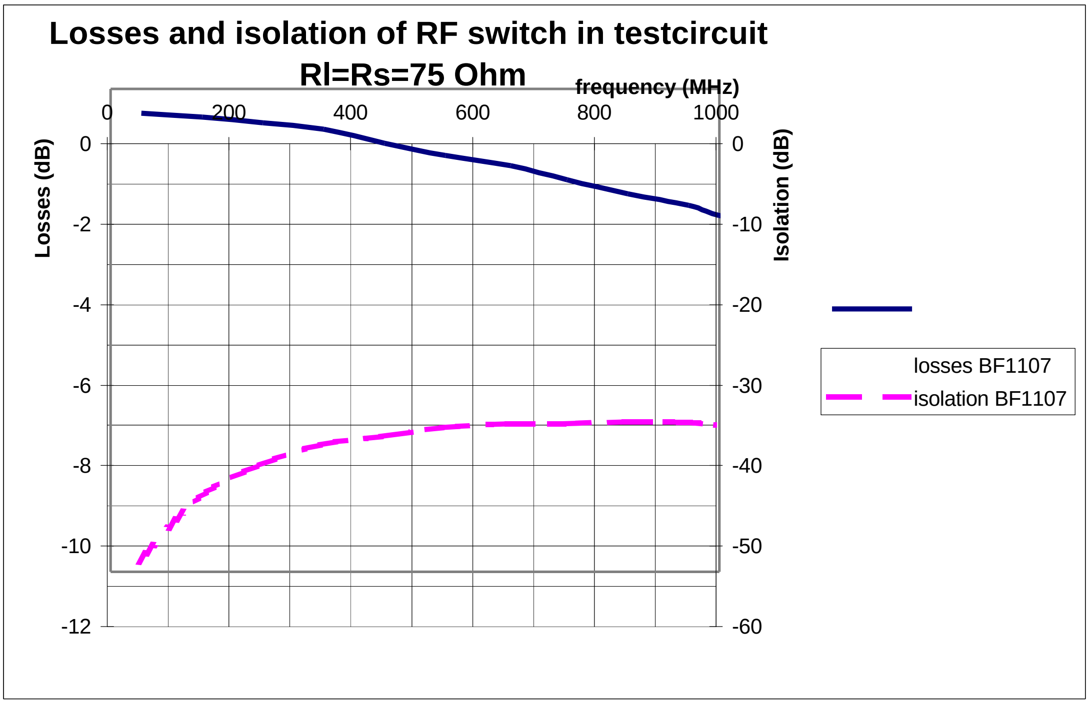


Fig. 6

The isolation (Mosfet is switched off) in the testcircuit is mainly determined by the feedback of the Mosfet in common Gate plus the parasitic capacitance of the testcircuit between Drain and Source. This parasitic capacitance must be very small.
The losses (Mosfet is switched on) in the testcircuit are at low frequencies determined by the $R_{DS\ on}$ of the Mosfet and at high frequencies by the $R_{DS\ on}$ and the Drain - Gate and Source - Gate capacitances of the Mosfet.
The parasitic capacitances of the circuit must be kept much lower than the capacitances of the Mosfet.

SPECIAL MEASURES TO BE TAKEN

In Fig. 4 only the principle of the application circuit of the switch in the VCR is given.
In the practical application circuit of a VCR the input and output of the wide band splitter amplifier are connected to the input and output of the switch.
As stated in chapter 3 the losses in the on situation of the switch are also determined by the capacitances at the input and the output of the switch.
If in the principle circuit of Fig.4 the Mosfet is switched on, then the wide band

splitter amplifier is still connected to the RF switch. This results into higher losses. Therefore special measures are needed to reduce the influence of the presence of the amplifier on the losses.
Theoretically this can be done by disconnecting the input as well as the output of the amplifier from the switch.
In practice this disconnecting can be done with a switch.
The principle of the circuit is then as given in Fig. 7.

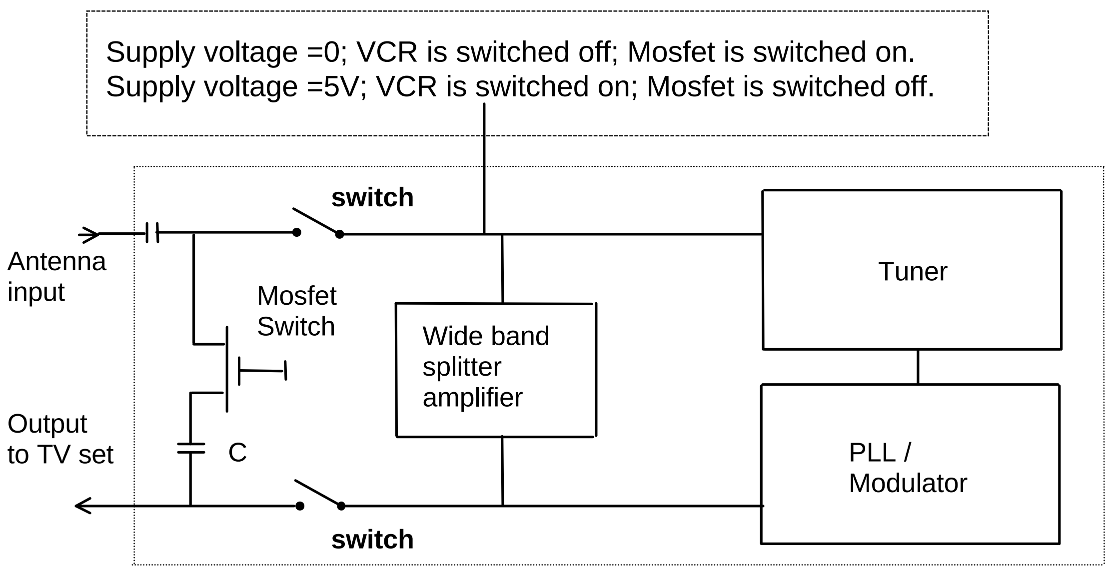


Fig. 7

The losses of the two switches in Fig. 7 must have low resistance if this switch is on and low capacitance if this switch is off. Such switches can be made with diodes. With the right choice of the diodes the resistance is low if the diode is forward biased and the capacitance is low if the bias voltage of the diode is 0V. Diodes that can be applied are bandswitching diodes (e.g. BA792 or BA277).
If the two stages of the wide band splitter amplifier are biased via the diode switches then the amplifier is “disconnected” from the switch if the supply voltage is 0V and “connected” if the supply voltage is 5V.
The main part of the circuit is then as given in Fig. 8 next page.

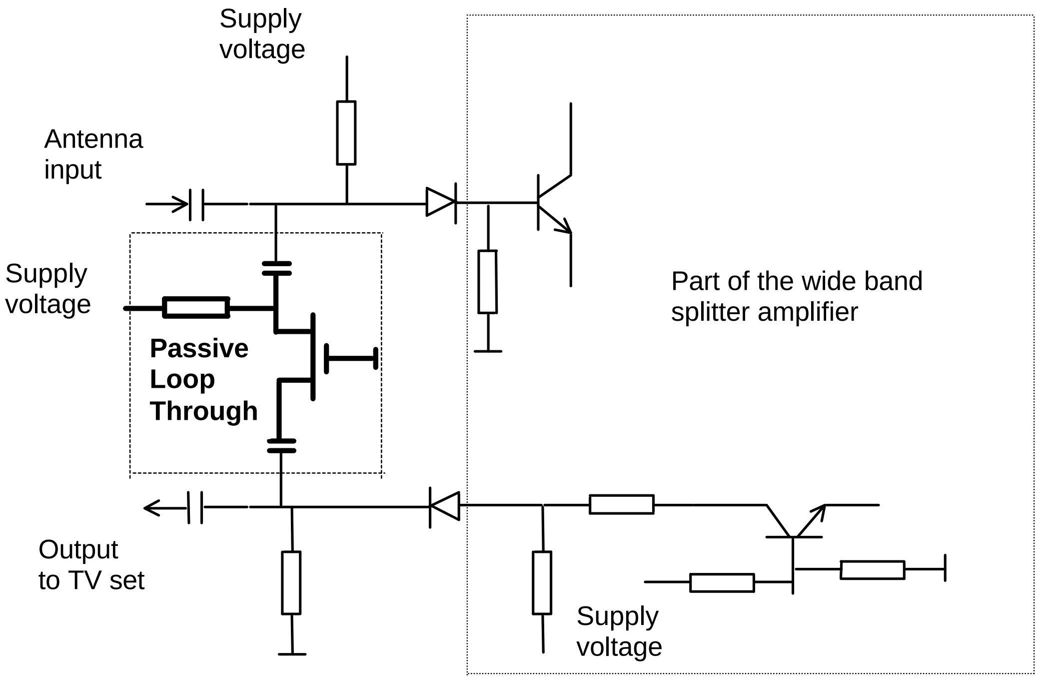


Fig. 8

CONCLUSIONS

The BF1107 is a specially developed triode Mosfet for the application of RF switch. In the on condition of the switch as well as in the off condition no D.C current flows through the Mosfet.
One of the application areas is the "Passive Loop Through" in a VCR.
The requirements for this application are:

Losses:	typ 2dB	max. 4dB.
Isolation:	> 30dB.	

This can be achieved with a BF1107 in the circuit of Fig. 8.
If this switch is applied the supply voltage of the VCR can be switched off in the "stand - by" condition of the VCR.
The R.F signal path to the T.V. set is then via the switch and not via a (power consuming) wide band splitter amplifier.

APPLICATION OF THE RF SWITCH BF1108

(BF1107 + BA277 in a SOT143 package)

INTRODUCTION

The BF1108 is a small signal RF switching Mosfet that can be used for switching RF signals up to 1GHz with good performance and switching RF signals up to 2GHz with reasonable performance. (See Fig. 1 for the circuit diagram).

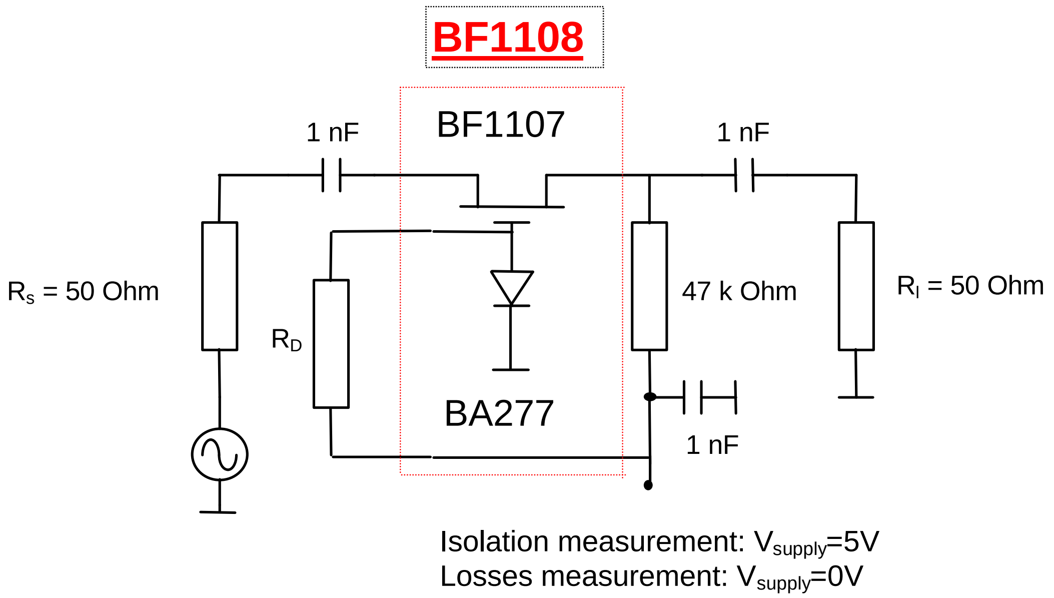


Fig. 1

The BF1108 consists of the RF switch BF1107 with a diode BA277 connected in series with the Gate. Drain and Source are interchangeable.
Both, the BF1107 and BA277 were mounted in one SOT143 package.

RF SWITCH BF1108 IN ITS APPLICATION

The losses of the RF switch are determined by the on resistance of the BF1107 and the capacitances at the input and the output to ground.
If no supply voltage is present at the RF switch (switch is on) than the gate of the BF1107 is connected to ground via the small capacitance of the diode BA277.
This will result in improved losses, especially at high frequencies. This is because input and output capacitance of the switch are lowered.
The isolation of the RF switch is determined by its' off resistance in parallel with the feedback capacitance and the impedance between gate and ground.
If there is a 5V supply voltage present at the switch (switch is off) than the gate of the BF1107 is connected to ground via the small seriesresistance of the BA277.
The impedance between gate and ground is mainly determined by the inductance

from gate to ground, especially at high frequencies. Therefore the small extra series resistance of the BA277 will have marginal influence on the isolation of the switch. However, the extra series inductance has influence on the isolation, especially at high frequencies.
For the BF1107 no current is needed as well in the on state as in the off state.
For the BF1108 also no current is needed for the on state. In the off state a small current through the BA277 is needed to ensure relatively low series resistance.

MEASUREMENTS ON THE BF1108

On the BF1108 we have measured the losses in the on state (V_{supply} = 0V) and the isolation in the off state (V_{supply} = 5V) in a 50Ω test circuit (see Fig. 1).
For comparison we have also measured the losses and the isolation of a BF1107.
The results of the measurements on a BF1107 are given in Graph. 1.
The results of the measurements on a BF1108 are given in the Graphs. 2 and 3.
In Graph 2 the results are given with a bias resistor (to the BA277) of 4.7kΩ.
This is a d.c. forward current through the diode of appr. 1mA.
Graph 3 shows the results with a d.c. current through the diode of appr. 2mA.
(Bias resistor to the diode 2.2kΩ).
In the specification the losses and the isolation are specified up to 860MHz.
However, it is possible to use the BF1108 also at higher frequencies with somewhat less performance. For information we have also measured the BF1108 and BF1107 at frequencies up to 2.05GHz.
The results of these measurements are given in Graph 4.

INFLUENCE OF PARASITIC CAPACITANCES

It is obvious that parasitic capacitances will influence the performance of the RF switch, also, additional feedback as well as additional parallel capacitances in parallel with the BF1108.
Measurements are done with additional parallel capacitances between Drain and Ground and between Source and Ground (see Fig. 2, next page).
We have also added some additional feedback between Drain and Source.
The results of these measurements are given in Graph 5.

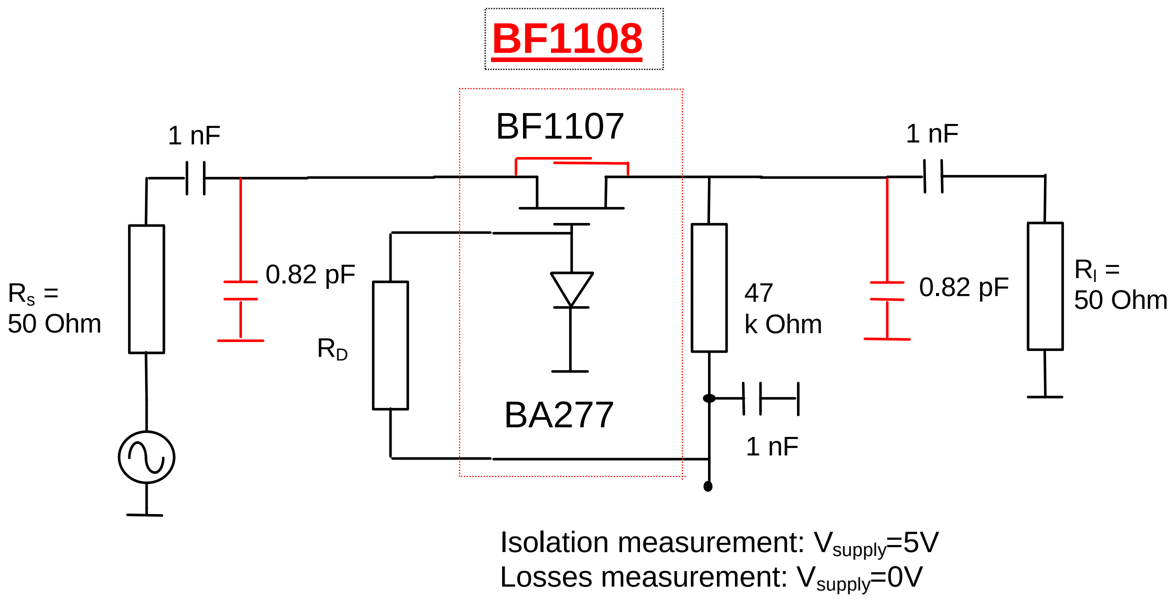
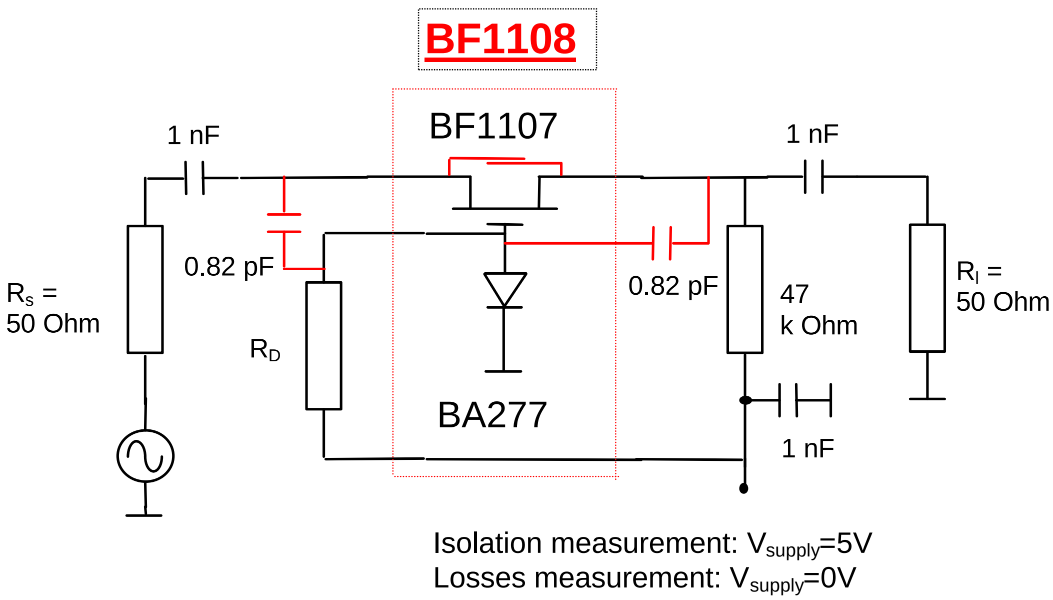


Fig. 2

The additional feedback was made by a short wire connected to the Drain, bending it towards the Source.

We have also done measurements with additional capacitances between Drain and Gate and between Source and Gate. Also with additional feedback between Drain and Source. Than the circuit diagram is as given in Fig. 3.



DISCUSSION ABOUT THE MEASURED RESULTS

Comparison of graphs 1, 2 and 3 show that at 1GHz the losses have been improved by appr. 0.5dB if a BF1108 is used i.s.o. a BF1107. Graph 4 shows that the BF1108 can also be used at frequencies higher than 1GHz with relatively reasonable performance. The losses at 2GHz are appr. 2.4dB for a BF1108 and > 5dB for a BF1107. At frequencies > 1GHz the isolation of a BF1108 is worse compared to that of the BF1107. This is caused by the seriesinductance of the BA277 (with bonding wires) to ground. If additional parallel capacitance is present at the input and the output of the BF1108 (Graph 5) the losses increase. We have done measurements with 0.82pF added. This increases the losses at 1GHz to appr. the same value as with the BF1107. This is because the advantage of the BF1108 with respect to the BF1107 is a reduction of the capacitance to ground if the switch is on and for these measurements we have increased this capacitance. The additional feedback capacitance results as can be expected in worse isolation and has almost no influence on the losses.
An explanation for this behaviour can be given with the help of the Figs. 4 and 5 which are simplified circuit diagrams of the BF1108 in the on state and off state respectively.

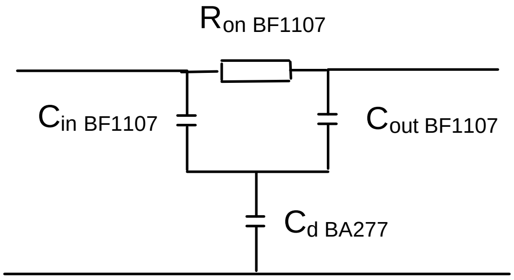


Fig. 4: Simplified circuit diagram of the BF1108 in the on state.
The losses in this circuit are mainly determined by the R_{on} of the BF1107, especially if the capacitance of the BA277 is small. If parallel capacitances at the input and output are present this will result in additional signal loss, especially at high frequencies. A small additional feedback capacitance in parallel with the relatively low ohmic R_{on} will have no influence on the losses.

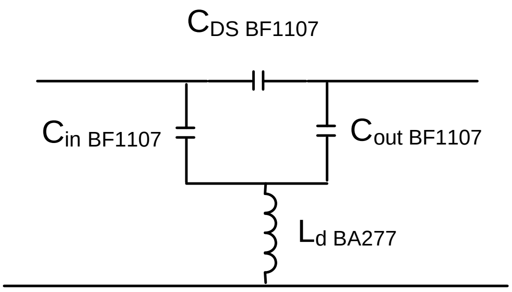


Fig. 5: Simplified circuit diagram of the BF1108 in the off state.
The isolation in this circuit is not only determined by the signal transfer via the feedback capacitance C_{DS} but also by the signal transfer via C_{in}, L_d and C_{out}.
These two kinds of signal transfer have (roughly) opposite phases.

If the signal transfer is only determined by the feedback capacitance one would expect a decrease in the isolation with 6dB / octave (if $1/\omega C_{DS} << (R_s + R_l)$). However, due to the opposite phases of the two signal transfers there will be a dip in the graph for the isolation as a function of frequency. In the Graphs 2, 3, 4 and 5 this dip is present at about 950MHz if no additional feedback is present. If additional feedback is present the dip shifts to a higher frequency. More series inductance in the Gate shifts the dip to a lower frequency. This can be seen from Graph 4 where the BF1107 and BF1108 are compared. For the BF1107 the dip is present at appr. 1850 MHz. The BF1108 shows a dip at appr. 950MHz. Due to this the isolation of the BF1108, compared to that of the BF1107, is worse at frequencies > 1GHz..

If additional capacitance (0.82pF) is present between Drain and Gate and Source and Gate, then the losses increase by appr. 0.25dB at 1GHz. This can also be explained from the fact that the capacitive signal path to ground is lower ohmic than without this additional capacitance. The influence of additional capacitances is much lower than connecting them directly to ground. This is because of the presence of the small diode capacitance. The isolation as a function of frequency is very dependent on the presence of the additional capacitances. (Compare Graph 3 and 6). We see that the dip in the curve is shifted to appr. 650MHz. And now the isolation at 1GHz is worse than 30dB. As stated before the dip can be shifted to a higher frequency if additional feedback is present. Than the dip can again be set to appr. 950MHz. The isolation at lower frequencies is than worsened, but now at 1GHz an isolation of > 30dB can be achieved (see Graph 6). Additional feedback does not influence the losses.

CONCLUSIONS

The BF1108 is an RF switch which has low losses at frequencies up to 2GHz. In the on state the losses are 1.15dB typical at 50MHz slowly increasing to 1.4dB typical at 1GHz and 2.4dB typical at 2GHz. The losses are strongly dependent on additional parallel capacitances present at the input and the output of the switch and almost not dependent on additional feedback between Drain and Source. The isolation of the BF1108 is > 50dB at 50MHz decreasing to appr. 35dB typical at 1GHz and appr. 15dB typical at 2GHz. The Graphs for the isolation as a function of frequency show a dip at a certain frequency. This dip is caused by a compensation effect of a signal transfer via the Drain - Source feedback capacitance and a signal transfer via the input- and output capacitances and the series inductance in the Gate. These two signal transfers have opposite phase which causes the dip in the curve. From this we can also conclude that the losses are strongly dependent on the feedback capacitance, the Drain - Gate and the Source - Gate capacitances and the series inductance in the Gate. Additional feedback capacitance shifts the dip in the curve isolation as a function of frequency to a higher frequency. The isolation at low frequencies is than worsened. Additional capacitances between Drain and Gate and Source and Gate and more series inductance in the Gate shifts the dip to lower frequencies.

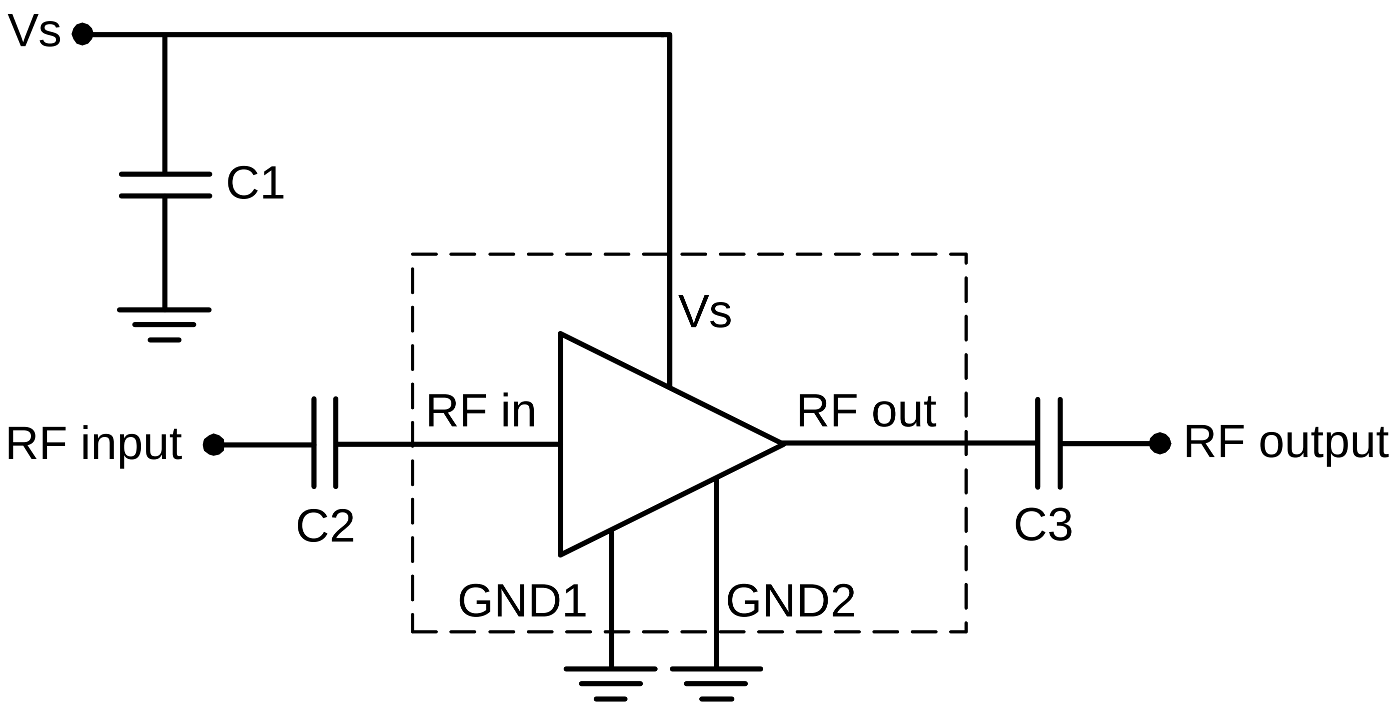
Appendix E: BGA2715-17 general purpose wideband amplifiers, 50 Ohm Gain Blocks

APPLICATION INFORMATION BGA2715-17

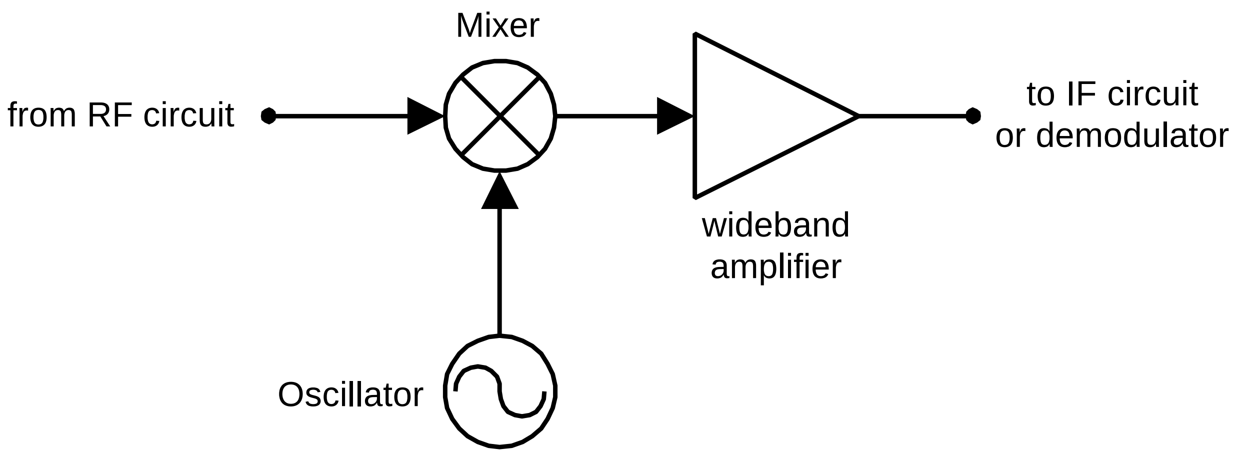
Figure 2 shows a typical application circuit for the BGA2715-17 MMIC.
The device is internally matched to 50 O, and therefore does not need any external matching. The value of the input and output DC blocking capacitors C2 and C3 should not be more than 100 pF for applications above 100 MHz. However, when the device is operated below 100 MHz, the capacitor value should be increased.

The 22 nF supply decoupling capacitor C1 should be located as close as possible to the MMIC.

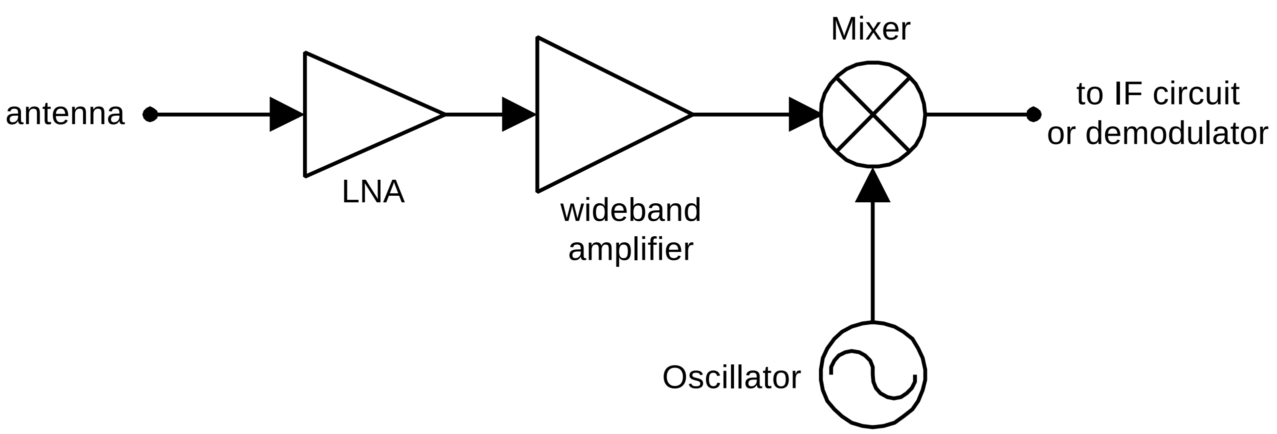
The PCB top ground plane, connected to the pins 2, 4 and 5 must be as close as possible to the MMIC, preferably also below the MMIC. When using via holes, use multiple via holes, as close as possible to the MMIC.



Application examples



The MMIC is very suitable as IF amplifier in e.g. LNB's. The exellent wideband characteristics make it an easy building block.



As second amplifier after an LNA, the MMIC offers an easy matching, low noise solution.

MMIC wideband amplifier

BGA2715

FEATURES

- Internally matched to 50 Ohms
- Wide frequency range, 3 dB bandwidth = 3.3 GHz
- Flat 22 dB gain, ± 1 dB up to 2.8 GHz
- -8 dBm output power at 1 dB compression point
- Good linearity for low current, OIP3 = 2 dBm
- Low second harmonic, -30 dBc at P_{Drive} = - 40 dBm
- Unconditionally stable, K

APPLICATIONS

- LNB IF amplifiers
- Cable systems
- ISM
- General purpose

DESCRIPTION

Silicon Monolitic Microwave Integrated Circuit (MMIC) wideband amplifier with internal matching circuit in a 6-pin SOT363 plastic SMD package.

PINNING

PIN	DESCRIPTION
1	V_S
2,5	GND 2
3	RF out
4	GND 1
6	RF in

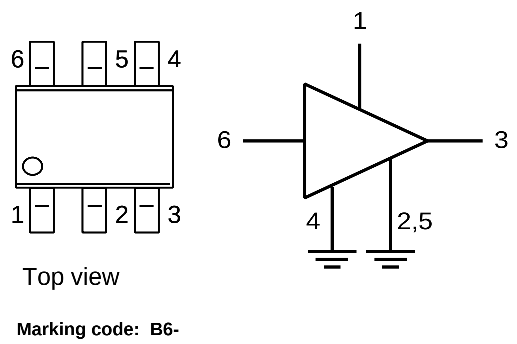


Fig.1 Simplified outline (SOT363) and symbol.

QUICK REFERENCE DATA

SYMBOL	PARAMETER	CONDITIONS	TYP.	MAX.	UNIT
Vs	DC supply voltage		5	6	V
Is	DC supply current		4.3	-	mA
$\|S21\|^2$	insertion power gain	f = 1 GHz	22	-	dB
NF	noise figure	f = 1 GHz	2.6	-	dB
$P_{L\ sat}$	saturated load power	f = 1 GHz	-4	-	dBm

MMIC wideband amplifier

BGA2716

FEATURES

- Internally matched to 50 Ohms
- Wide frequency range, 3 dB bandwidth = 3.2 GHz
- Flat 23 dB gain, ± 1 dB up to 2.7 GHz
- 9 dBm output power at 1 dB compression point
- Good linearity for low current, OIP3 = 22 dBm
- Low second harmonic, -38 dBc at P_{Load} = - 5 dBm
- Unconditionally stable, K > 1.2

APPLICATIONS

- LNB IF amplifiers
- Cable systems
- ISM
- General purpose

DESCRIPTION

Silicon Monolitic Microwave Integrated Circuit (MMIC) wideband amplifier with internal matching circuit in a 6-pin SOT363 plastic SMD package.

PINNING

PIN	DESCRIPTION
1	V_S
2,5	GND 2
3	RF out
4	GND 1
6	RF in

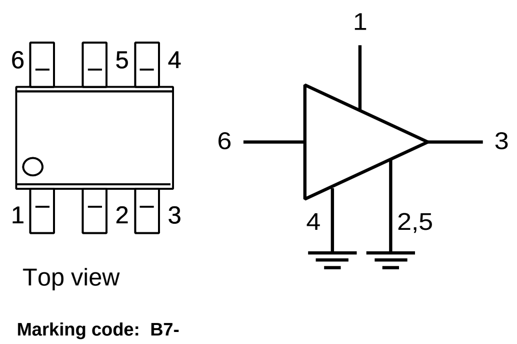


Fig.1 Simplified outline (SOT363) and symbol.

QUICK REFERENCE DATA

SYMBOL	PARAMETER	CONDITIONS	TYP.	MAX.	UNIT
Vs	DC supply voltage		5	6	V
Is	DC supply current		15.9	-	mA
$\|S21\|^2$	insertion power gain	f = 1 GHz	22.9	-	dB
NF	noise figure	f = 1 GHz	5.3	-	dB
$P_{L\ sat}$	saturated load power	f = 1 GHz	11.6	-	dBm

MMIC wideband amplifier

BGA2717

FEATURES

- Internally matched to 50 Ohms
- Wide frequency range, 3 dB bandwidth = 3.2 GHz
- Flat 24 dB gain, ± 1 dB up to 2.8 GHz
- -2.5 dBm output power at 1 dB compression point
- Good linearity for low current, OIP3 = 10 dBm
- Low second harmonic, -38 dBc at P_{Drive} = - 40 dBm
- Low noise figure, 2.3 dB at 1 GHz.
- Unconditionally stable, K > 1.5

APPLICATIONS

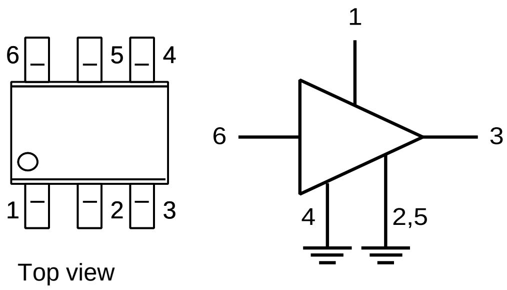
- LNB IF amplifiers
- Cable systems
- ISM
- General purpose

DESCRIPTION

Silicon Monolitic Microwave Integrated Circuit (MMIC) wideband amplifier with internal matching circuit in a 6-pin SOT363 plastic SMD package.

PINNING

PIN	DESCRIPTION
1	V_S
2,5	GND 2
3	RF out
4	GND 1
6	RF in



Marking code: 1B-

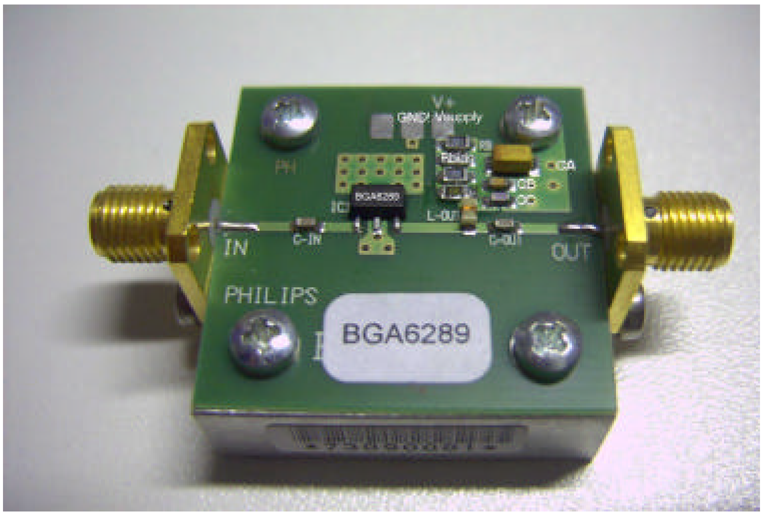
Fig.1 Simplified outline (SOT363) and symbol.

QUICK REFERENCE DATA

SYMBOL	PARAMETER	CONDITIONS	TYP.	MAX.	UNIT
Vs	DC supply voltage		5	6	V
Is	DC supply current		8.0	-	mA
$\|S21\|^2$	insertion power gain	f = 1 GHz	24	-	dB
NF	noise figure	f = 1 GHz	2.3	-	dB
$P_{L\ sat}$	saturated load power	f = 1 GHz	1	-	dBm

Appendix F: BGA6x89 general purpose medium power amplifiers, 50 Ohm Gain Blocks

Application note for the BGA6289



Application note for the BGA6289.
(See also the objective datasheet BGA6289)

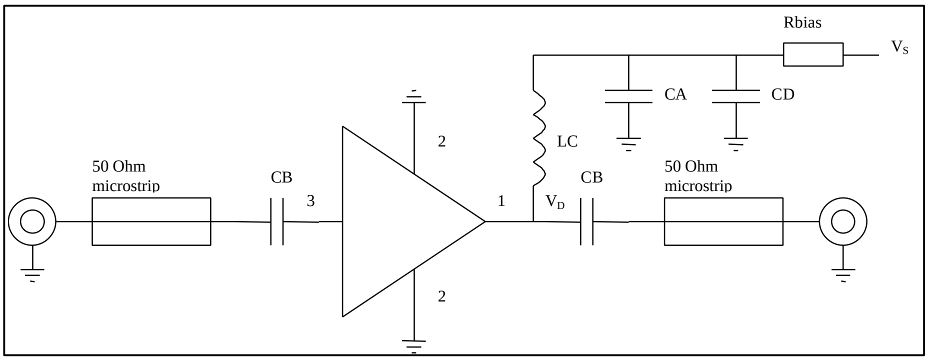


Figure 1 Application circuit.

COMPONENT	DESCRIPTION	VALUE	DIMENSIONS
C_{in} C_{out}	multilayer ceramic chip capacitor	68 pF	0603
C_A	Capacitor	1 µF	0603
C_B	multilayer ceramic chip capacitor	1 nF	0603
C_C	multilayer ceramic chip capacitor	22 pF	0603
L_{out}	SMD inductor	22 nH	0603
Vsupply	Supply voltage	6 V	
R_{bias}=RB	SMD resistor 0.5W	27 Ohm	----

Table 1 component values placed on the demo board.

C_A is needed for optimal supply decoupling .
Depending on frequency of operation the values of C_{in} C_{out} and L_{out} can be changed (see table 2).

COMPONENT	**Frequency (MHz)**				
	500	800	1950	2400	3500
C_{in} C_{out}	220 pF	100 pF	68 pF	56 pF	39 pF
C_A	1 µF	1 µF	1 µF	1 µF	1 µF
C_B	1 nF	1 nF	1 nF	1 nF	1 nF
C_C	100 pF	68 pF	22 pF	22 pF	15 pF
L_{out}	68 nH	33 nH	22 nH	18 nH	15 nH

Table 2 component selection for different frequencies.

V_{supply} depends on R_{bias} used. Device voltage must be approximately 4 V (i.e. device current = 80mA).

With formula 1 it is possible to operate the device under different supply voltages.
If the temperature raises the device will draw more current, the voltage drop over Rbias will increase and the device voltage decrease, this mechanism provides DC stability.
Measured small signal performance.

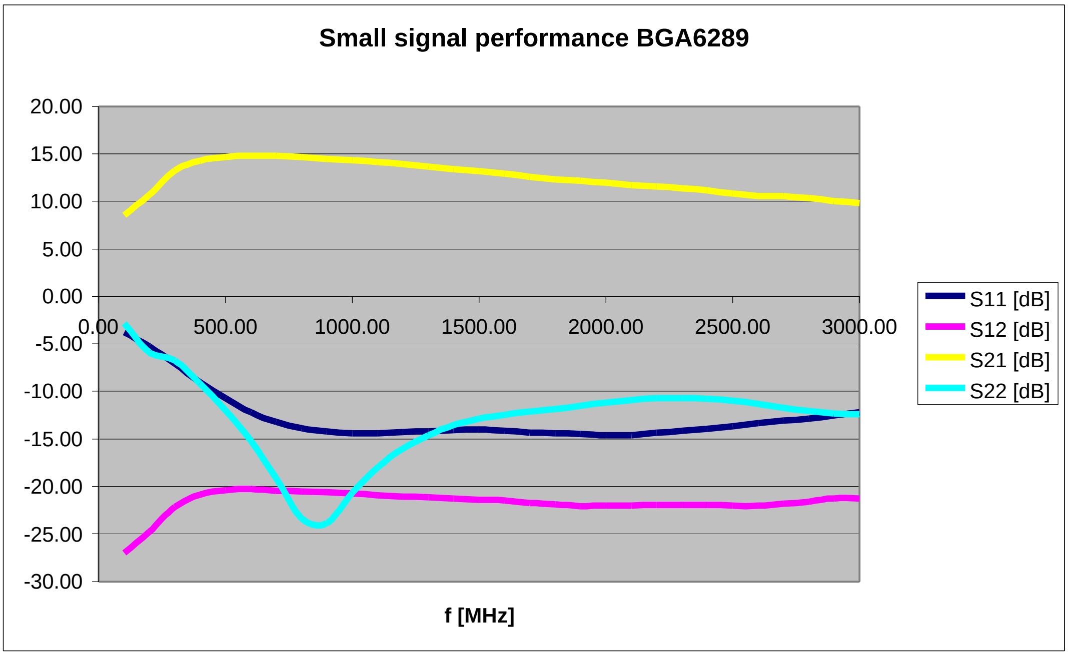


Figure 2 Small signal performance.

Measured large signal performance.

f	850 MHz	2500 MHz
$IP3_{out}$	31 dBm	25 dBm
PL_{1dB}	18 dBm	16 dBm
NF	3.8	4.1

Table 3 Large signal performance and noise figure.

Application note for the BGA6489

Application note for the BGA6489.
(See also the objective datasheet BGA6489)

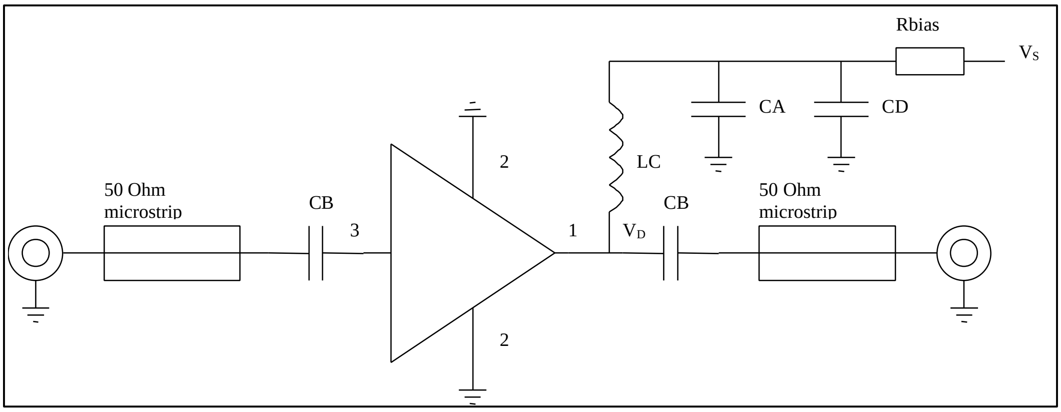


Figure 1 Application circuit.

COMPONENT	DESCRIPTION	VALUE	DIMENSIONS
C_{in} C_{out}	multilayer ceramic chip capacitor	68 pF	0603
C_A	Capacitor	1 µF	0603
C_B	multilayer ceramic chip capacitor	1 nF	0603
C_C	multilayer ceramic chip capacitor	22 pF	0603
L_{out}	SMD inductor	22 nH	0603
Vsupply	Supply voltage	8 V	
R_{bias}=RB	SMD resistor 0.5W	33 Ohm	----

Table 1 component values placed on the demo board.

C_A is needed for optimal supply decoupling .
Depending on frequency of operation the values of C_{in} C_{out} and L_{out} can be changed (see table 2).

COMPONENT	Frequency (MHz)				
	500	800	1950	2400	3500
C_{in} C_{out}	220 pF	100 pF	68 pF	56 pF	39 pF
C_A	1 µF	1 µF	1 µF	1 µF	1 µF
C_B	1 nF	1 nF	1 nF	1 nF	1 nF
C_C	100 pF	68 pF	22 pF	22 pF	15 pF
L_{out}	68 nH	33 nH	22 nH	18 nH	15 nH

Table 2 component selection for different frequencies.

V_{supply} depends on R_{bias} used. Device voltage must be approximately 5.1 V (i.e. device current = 80mA).

With formula 1 it is possible to operate the device under different supply voltages.

If the temperature raises the device will draw more current, the voltage drop over Rbias will increase and the device voltage decrease, this mechanism provides DC stability.
Measured small signal performance.

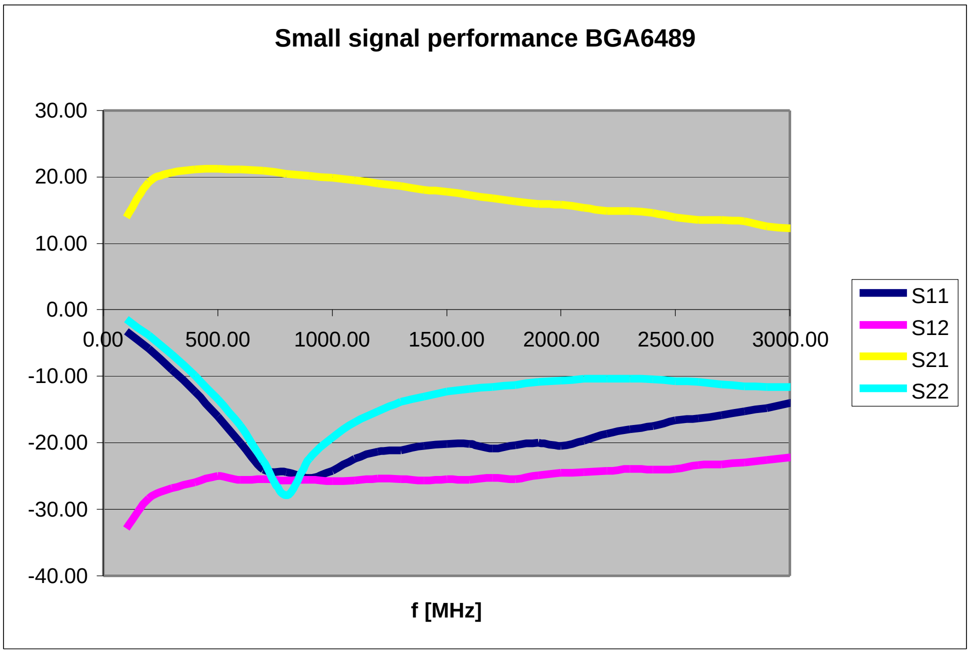


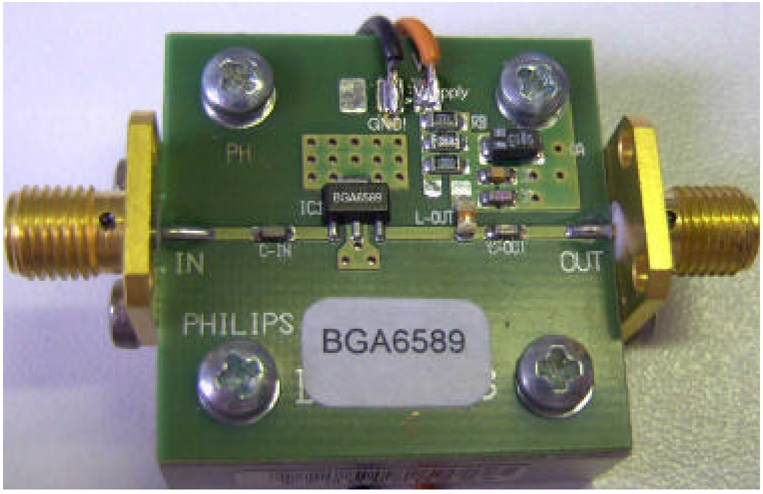
Figure 2 Small signal performance.

Measured large signal performance.

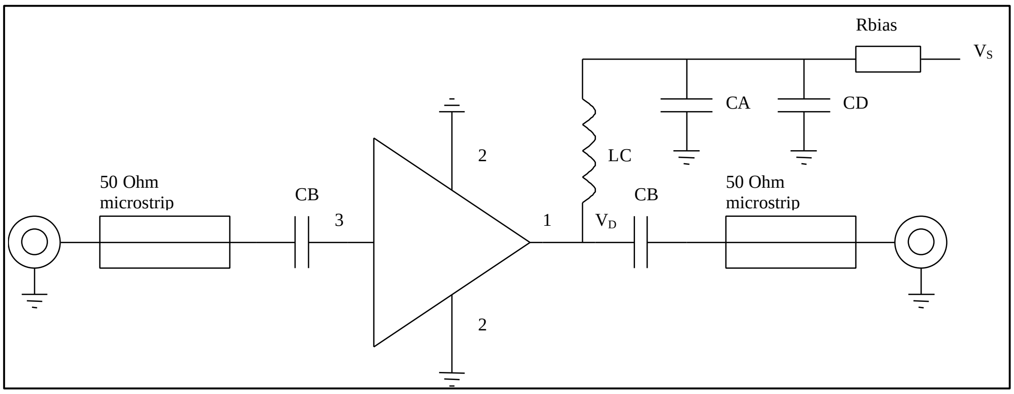
f	850 MHz	2500 MHz
$IP3_{out}$	33 dBm	27 dBm
PL_{1dB}	20 dBm	17 dBm
NF	3.1 dB	3.4 dB

Table 3 Large signal performance and noise figure.

Application note for the BGA6589



The Demo Board with medium power wide-band gainblock BGA6589.
(See also the objective datasheet BGA6589)



Application circuit.

COMPONENT	DESCRIPTION	VALUE	DIMENSIONS
C_{in} C_{out}	multilayer ceramic chip capacitor	68 pF	0603
CA	Capacitor	1 µF	
C_B	multilayer ceramic chip capacitor	1 nF	0603
C_C	multilayer ceramic chip capacitor	22 pF	0603
L_C	SMD inductor	22 nH	0603
Vsupply	Supply voltage	7.5 V	
R_{bias}=RB	SMD resistor 0.5W	33 Ohm	----

Table 1 component values placed on the demo board.

C_A is needed for optimal supply decoupling .
Depending on frequency of operation the values of C_{in} C_{out} and L_{out} can be changed (see table 2).

COMPONENT	Frequency (MHz)				
	500	800	1950	2400	3500
C_{in} C_{out}	220 pF	100 pF	68 pF	56 pF	39 pF
C_A	1 µF	1 µF	1 µF	1 µF	1 µF
C_B	1 nF	1 nF	1 nF	1 nF	1 nF
C_C	100 pF	68 pF	22 pF	22 pF	15 pF
L_{out}	68 nH	33 nH	22 nH	18 nH	15 nH

Table 2 component selection for different frequencies.
V_{supply} depends on R_{bias} used. Device voltage must be approximately 4.8 V (i.e. device current = 83mA).

With formula 1 it is possible to operate the device under different supply voltages.
If the temperature raises the device will draw more current, the voltage drop over Rbias will increase and the device voltage decrease, this mechanism provides DC stability.
Measured small signal performance.

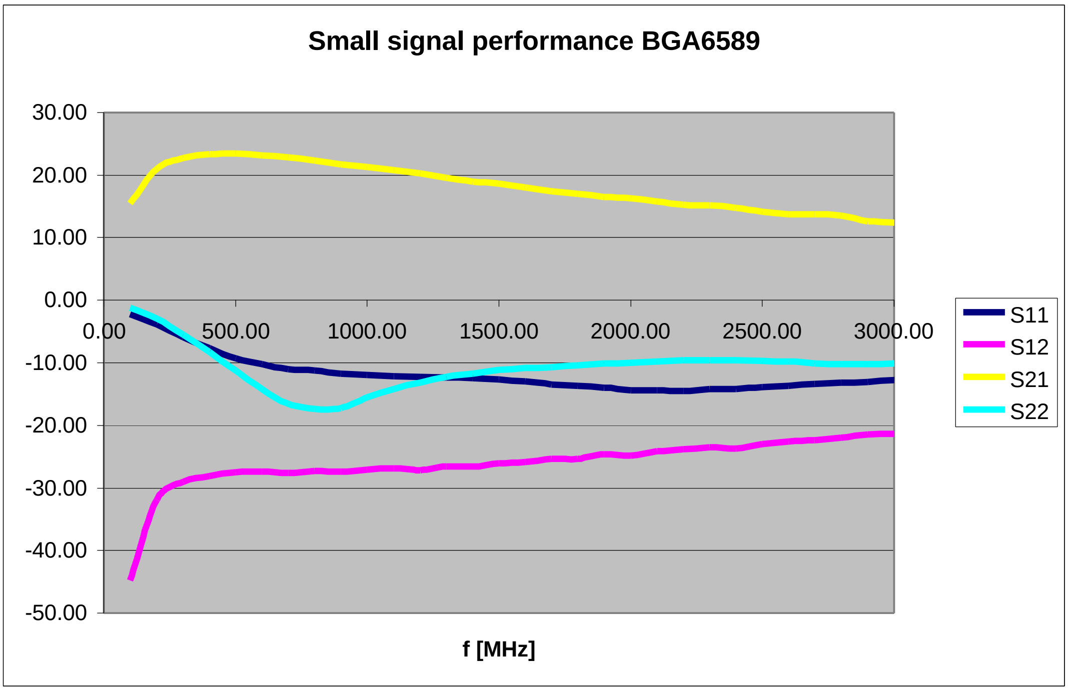


Figure 2 Small signal performance.

Measured large signal performance.

f	850 MHz	2500 MHz
$IP3_{out}$	33 dBm	32 dBm
PL_{1dB}	21 dBm	19 dBm
NF	3.1 dB	3.4 dB

Table 3 Large signal performance and noise figure.