Misure Elettroniche - Programma svolto nell'A.A. 2009-2010

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- Richiami e definizioni. Decibel e unità logaritmiche assolute. Attenuatori resistivi. Serie di Fourier, serie di Fourier di onda quadra. Rumore termico, cenno a rumore granulare e rumore di contatto 1/f. Densità di potenza di rumore e sua misura, banda equivalente di rumore. Guadagno di potenza e funzione di trasferimento di un quadripolo. Fattore e cifra di rumore. Cifra di rumore della cascata di N quadripoli non interagenti, discussione e semplici applicazioni. Linee di trasmissione. Caso di linea corta rispetto alla lunghezza d'onda chiusa in cc oppure in ca. Caso generale (sempre per $l<<\lambda$). Analisi della cascata infinita di celle L-C. Impedenza e funzione di trasferimento. Dalle costanti concentrate alle costanti distribuite. Caso di linea infinita o adattata. Onda diretta, espressione matematica (fasori e dominio del tempo in regime sinusoidale). Lunghezza d'onda, velocità di propagazione, verso di propagazione. Relazione fra parametri per unità di lunghezza, impedenza caratteristica, velocità di propagazione. Equazioni per tensione e corrente lungo la linea di trasmissione, onda diretta e onda inversa, coefficiente di riflessione (lungo la linea, al carico, al generatore). Rapporto d'onda stazionaria. Impedenza (lungo la linea e all'ingresso). Espressione esplicita di tensione e corrente. Analisi con le riflessioni multiple e interpretazione fisica del risultato dell'analisi. Potenza che fluisce lungo la linea. Potenza diretta e inversa, potenza disponibile. Perdite: come si modificano le equazioni fondamentali per includere le perdite. Attenuazione nei cavi. Casi particolari (generatore e/o carico adattati). Comportamento reale dei componenti. Resistore. Condensatore (inclusa derivazione induttanza parassita interna per condensatore con armature piane parallele e circolari). Induttore (con discussione del fenomeno del rilassamento e perdite per isteresi del materiale magnetico). Induttori mutuamente accoppiati. Corto-circuito (inclusa trattazione dell'induttanza interna per filo tondo e formule per cavo coassiale). Circuito aperto.

- Analizzatore di spettro. Presentazione del visore (REF_LEV, SPAN, dB/, CF, ATT, RBW. Ricevitore supereterodina, oscillatore locale, segnale, frequenza intermedia. Mixer: sovrapposizione del segnale e dell'oscillatore locale, ipotesi $V_o>>V_s$ (prodotti di mescolazione esclusi). Non linearità intrinseca del mixer e prodotti di mescolazione $nf_o \pm f_s$ (nfo±mfs se Vs diviene comparabile con V_o). Equazione di sintonia: $f_o-f_s=f_{IF}$. Rivelatore di inviluppo a valle di filtro + amplificatore a frequenza intermedia. Necessità di bande di risoluzione variabili per risolvere componenti spettrali vicine ($\Delta f \approx B_3$). Perdita di conversione del mixer. Analisi della risposta del filtro IF (filtro gaussiano) ad un segnale con frequenza modulata linearmente. Condizione $T > k \cdot F/B_3^2$. Esempi numerici. Criteri progettuali analizzatore di spettro ($f_T<f_{LO,min}<f_{IF}$, frequenza di taglio inferiore). Livelli caratteristici: sensibilità, livello esente da spurie, punto di compressione ad 1 dB, danneggiamento mixer, danneggiamento attenuatore. Gamma dinamica massima e gamma dinamica istantanea. Tecnica per la verifica di eventuale comportamento non-lineare. Cifra di rumore dell'analizzatore di spettro. Rumore rivelato e densità di probabilità di Rayleigh. Filtro video e condizione $T > k \cdot F/(B_3 B_V)$. Valore medio del rumore sul display: scala lineare e scala logaritmica. Vari metodi di determinazione della cifra di rumore dell'analizzatore di spettro: 1) dal livello di rumore medio in scala lineare, 2) dal livello di rumore medio in scala logaritmica, 3) dal valore efficace della tensione rivelata, 4) da valutazione di un fattore di cresta. Tracking generator e misura della perdita di inserzione, relazione con funzione di trasferimento. Normalizzazione delle tracce. Risposta dell'analizzatore di spettro agli impulsi. Banda equivalente impulsiva. Casi in cui la frequenza di ripetizione degli impulsi è maggiore o minore della banda di risoluzione.

- Impedenzimetro vettoriale (HP 4193A, 400 kHz – 100 MHz): schema a blocchi e principio di funzionamento (con cenno a schema di principio di rivelatore di fase e traslazione di frequenza per tramite di campionamento). Parassiti (residenza fisica, valutazione ordine di grandezza, misura, modello circuitale, depurazione della misura dagli effetti dei parassiti). Analisi specifica nel caso del parassita ohmico-induttivo. Specifiche dello strumento, accuratezza, ripercussione degli errori di misura di ampiezza e fase su parte reale e parte immaginaria dell'impedenza, esempi.

- Wattmetro bolometrico (HP 432A, 10 MHz – 18 GHz): schema a blocchi semplificato e principio di funzionamento (basato su impiego di termistori). Dettaglio dello schema dello strumento di misura e della testa bolometrica HP 478A, 10 MHz – 10 GHz (in particolare evidenziata la separazione del circuito di polarizzazione da quello RF). Cenno alla necessità di mantenere la potenza dissipata sul termistore costante al variare della frequenza (DC vs. RF, considerazioni su effetto pelle e parassiti). Specifiche. Misure di precisione con wattmetro bolometrico: effetto del disadattamento del wattmetro e del generatore (fattore di calibrazione, correzione per riflessioni multiple).

- Cenno al principio di funzionamento dell'accoppiatore direzionale (linee accoppiate in mezzo omogeneo).

Simulation and automatic code generation for real time embedded systems using ScicosLab and Erika/Linux

Tutors:

Simone Mannori - ScicosLab developer (www.scicos.org, www.scicoslab.org)
Paolo Gai - Evidence S.r.l.
Claudio Macchiavelli - I.A.F. (Industrial Automation Freedom, Lugano, CH) (http://www.iafreed.com/)
Matteo Morelli - RTSS developer (Robotics Group at Centro "E. Piaggio", University of Pisa. http://rtss.sourceforge.net/)
Roberto Bucher - SUPSI Lugano

Location

Florence University - Plesso didattico, Viale Morgagni 40, Firenze, ITALY

Time

3 days (3 day course)
19/20/21 October 2010
Morning : 9:00 - 13:00
Afternoon : 14:00 - 17:00
Open Talk : 17:00 - 19:00

Web page:
http://erika.tuxfamily.org/scilabscicos/scicoslabcourse2010.html

Registration
Registration is mandatory:
https://www.softconf.com/b/scicoslab2010/cgi-bin/scmd.cgi?scmd=people&pageid=0

Languages
- Italian as default spoken languages, English and French on request; books and documents in English.

Target audience
- Students, engineers and scientists working with complex simulations and control systems.

Prerequisites
- A personal computer (Linux/ Windows/ Mac OSx)
- A clean USB key for file exchange
- Internet access is not required *but could be useful*

Note for Windows users: some exercises require the presence of a C compiler. Under Linux "gcc" is installed as default (check with "gcc -v"). Windows users must install Visual C/C++ Studio Express 2008 (this version is freely downloadable from Microsoft's web site).

References
- The new Yellow Book: Modeling and Simulation in Scilab/ Scicos With Scicoslab 4.4, Stephen L. Campbell, Jean-Philippe Chancelier, et Ramine Nikoukhah
- On line documentation available here: http://www.scicos.org/documentations.html

Rationale
This three days course is a general purpose introduction to the art of dynamical systems simulation and automatic code generation for real time embedded systems using ScicosLab/Scicos. We will make references and comparisons with Matlab/Simulink, Kepler and Ptolemy. The course is focused on ScicosLab/Scicos, ERIKA and Scicos-FLEX.

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Day One: "Basic".

Morning sessions
1./ **"ScicosLab/Scicos for dummies like us" -** Simone Mannori (2h)
- Gentle introduction to simulation and automatic code generation
- ScicosLab basic features
- What you can simulate with ScicosLab?
- Scicos: a dynamical systems simulator
- What a dynamical system is ?
- Mathematical models
- Differential equations (explicit and implicit form, linear and non linear)
- Modelica language and simulations
- Continuous, discrete and hybrid systems
- How Scicos works (basic introduction to the internal structure)
- Basic example of code generation

2./ **"Real Time Operating Systems for embedded applications"** Paolo Gai (2h)
- Introduction to embedded real-time systems

- Introduction to the FLEX Boards

- Introduction to the OSEK/VDX operating system and ERIKA Enterprise

Afternoon: workshops
1./ Simone Mannori
- How to configure and compile ScicosLab
- Familiarisation with ScicosLab
- Familiarisation with Scicos
- How to build a diagram and run a simulation
- Import/export simulation's results

Exercises (physical modeling, simulation, controller design)
- DC motor
- Three pole RC filter
- Coupled mechanical oscillator

Basic block development.

2./ Paolo Gai (1h)
- OSEK/VDX and ERIKA Enterprise Examples

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Day two:
"We do ScicosLab - and other things - not because they are easy but because they are hard".

Morning sessions
1./ "Scicos guts. No guts, no glory." - Simone Mannori (2h)
- ScicosLab from a development standpoint
- How to develop a simple Scicos block from scratch
- Interfacing function structure
- Computational function structure
- Examples of Scicos block development
- Scicos-FLEX: the Flexible Code Generator from Switzerland (Roberto Bucher)
- Scicos-FLEX for dummies
- Scicos-FLEX internals

2./ "Real Time Operating Systems for embedded applications" Paolo Gai (2h)
- OSEK/VDX and ERIKA Enterprise examples

- Details on the code generation using ScicosLab and the FLEX Boards

Afternoon: workshops
1./ Simone Mannori
- Scicos Blocks development with examples
- Scicos FLEX code generator usage and development (with examples)

2./ Paolo Gai (0,5-1h)

- Code generation examples with ScicosLab

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Day three: "ScicosLab is who ScicosLab does".

Morning sessions

1./ Real time communications: CAN and Powerlink - Claudio Macchiavelli (1.5h)
- Earlier communication between Control and Sensor/actuator
- Up-to-Date realtime digital communication overview
- Network time and the collision problem.
- Protocol type: Master/Slave or Distributed ?
- The real world:
  - Netwok: RS485, CAN, Ethernet, WLAN
  - Protocol: canOpen, Powerlink, Ethercat
- IEEE1588:A Precision Clock Synchronization Protocol
- The future

2./ ScicosLab/Scicos for robotics applications - Matteo Morelli (1.5h)
- RTSS: the Robotics Toolbox for Scilab/Scicos
- Rigid motions representation in R^3 with ScicosLab
- Robotic manipulators modelling with ScicosLab
- Robotic control systems simulation with Scicos
- Robot control code generation for use with Linux RTAI

3./ Wrap up session - Simone Mannori, Paolo Gai - (1h)
- What you have learnt during the last three days?
- What you have enjoyed/hated?
- Suggestion to improve the training course.

Afternoon: workshops

1./ Claudio Macchiavelli
- CAN in practice with IAF servo drives and ScicosLab based host station

2./ Matteo Morelli
- Installation of RTSS
- Review of some basic examples included in the distribution
- Kinematic models of simple planar manipulators from real data
- From task space to joint space references via kinematic inversion
- Adding dynamics to the models: dealing with real inertial and actuator/transmission parameters
- Joint space centralized control of the planar manipulators developed above
- Controller code generation for Linux RTAI

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IL DECIBEL (1 di 2)

Il decibel (si abbrevia dB) è definito da

$$10 \log \left(\frac{P_1}{P_2} \right)$$

dove P_1 e P_2 sono due **POTENZE** e il logaritmo è in base 10.

L'utilità del decibel risiede nel fatto che

1 - Il logaritmo comprime i valori alti ed espande i valori bassi $\Rightarrow$ il dB viene usato quando si ha a che fare con grandezze caratterizzate da una *grande dinamica*;

2 - Prodotto, divisione ed elevazione a potenza (operazioni molto comuni nella tecnica) vengono trasformati, usando i dB, in operazioni più semplici: somma, sottrazione e prodotto, rispettivamente.

Se P_1 e P_2 sono potenze dissipate sulla stessa resistenza R allora:

$$10 \log \left(\frac{P_1}{P_2} \right) = 10 \log \left(\frac{V_1^2/R}{V_2^2/R} \right) = 20 \log \left(\frac{V_1}{V_2} \right)$$

$$10 \log \left(\frac{P_1}{P_2} \right) = 10 \log \left(\frac{I_1^2 R}{I_2^2 R} \right) = 20 \log \left(\frac{I_1}{I_2} \right)$$

Anche il rapporto di tensioni e correnti può essere espresso in dB (usando il fattore 20 anziché 10 a moltiplicare) ma *solo se* tensioni e correnti corrispondono a potenze dissipate sulla stessa resistenza R.

Nella tecnica con il dB si usa esprimere anche il valore di grandezze assolute.

$\boxed{\text{dBm}}$ $10 \log \left(\frac{P}{1\ \text{mW}}\right)$, dove la potenza P è espressa in mW

$\boxed{\text{dB}(\mu\text{V})}$ $20 \log \left(\frac{V}{1\ \mu\text{V}}\right)$, dove la tensione V è espressa in μV

Se P è la potenza dissipata e V la tensione applicata alla stessa resistenza $R = 50\ \Omega$ allora

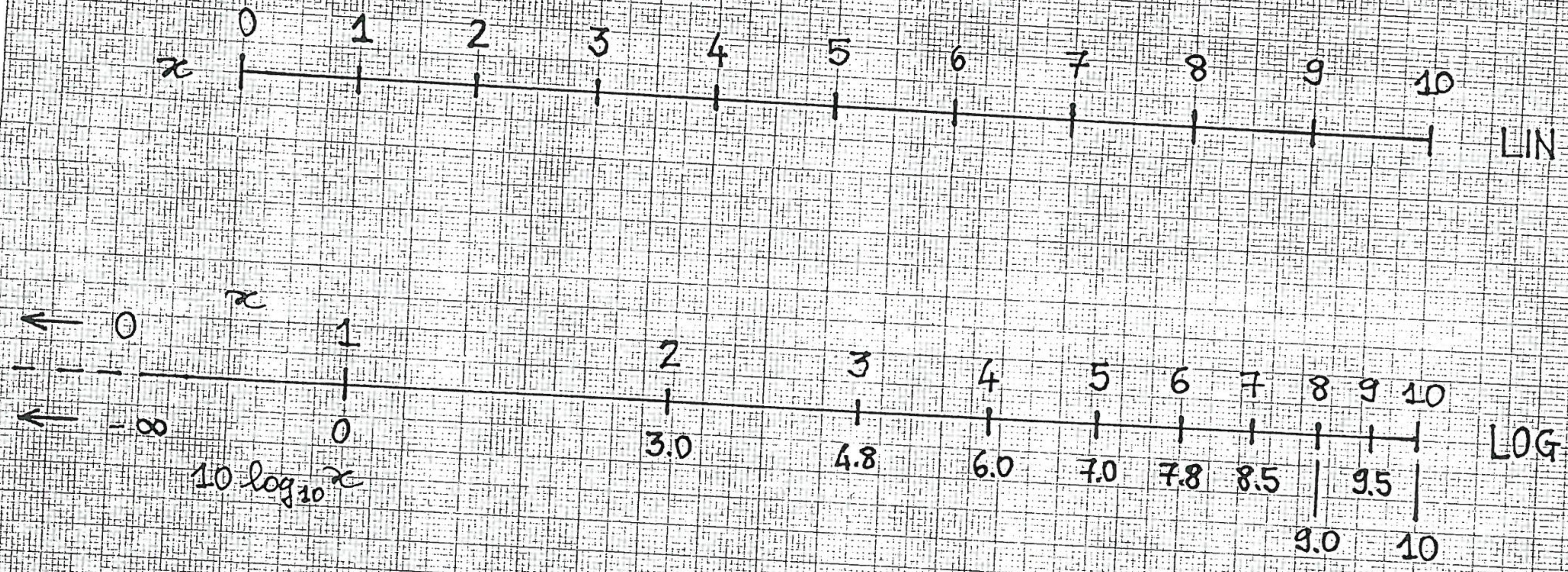
$$P(\text{in dBm}) = V(\text{in dB}(\mu\text{V})) - 107$$

Seguendo il principio su cui è basata la definizione di dBm e dB(μV) si possono definire dB(W/m^2), dB(μV/m), dB(μT), dB(...).

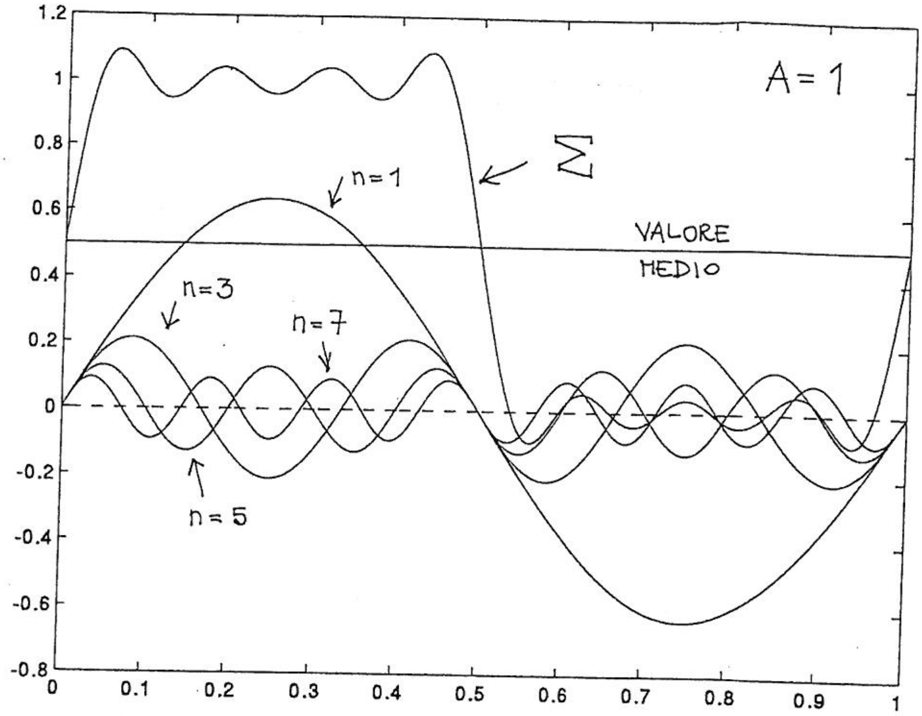
Tabella dei logaritmi

P_1/P_2	$10 \log(P_1/P_2)$
1	0
2	3.0
3	$\sim$ 4.75
4	6.0
5	7.0
6	$\sim$ 7.75
7	8.5
8	9.0
9	9.5

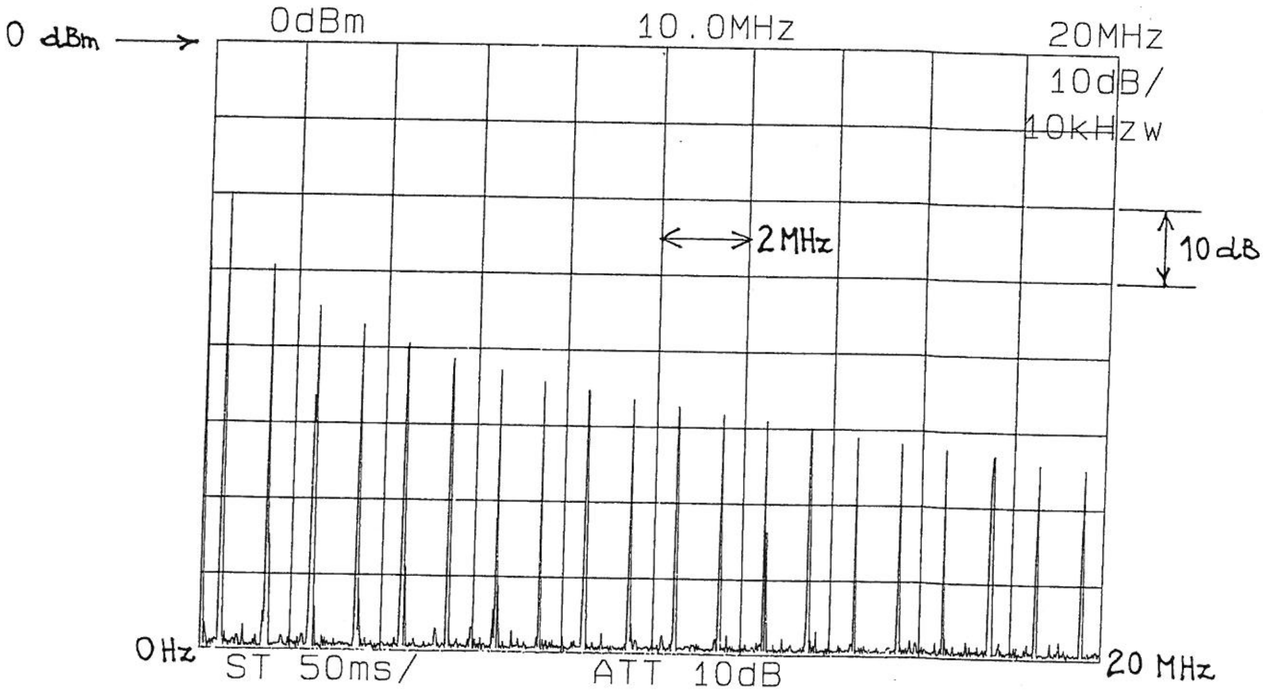
SCALA LINEARE (LIN) — SCALA LOGARITMICA (LOG)

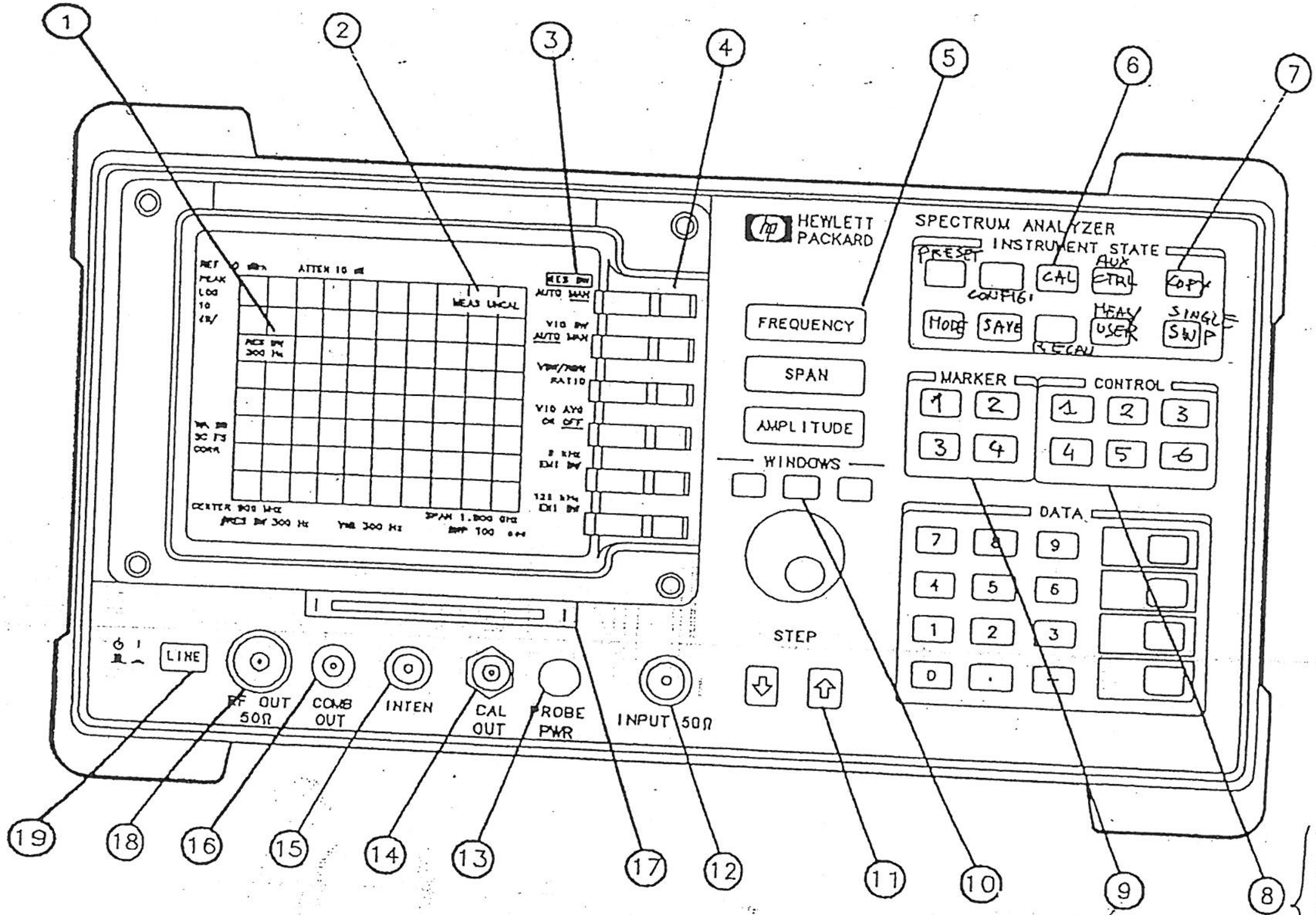


SERIE DI FOURIER DELL' ONDA QUADRA



SPETTRO ONDA QUADRA
AMPIEZZA 46 mV, FREQUENZA 500 kHz





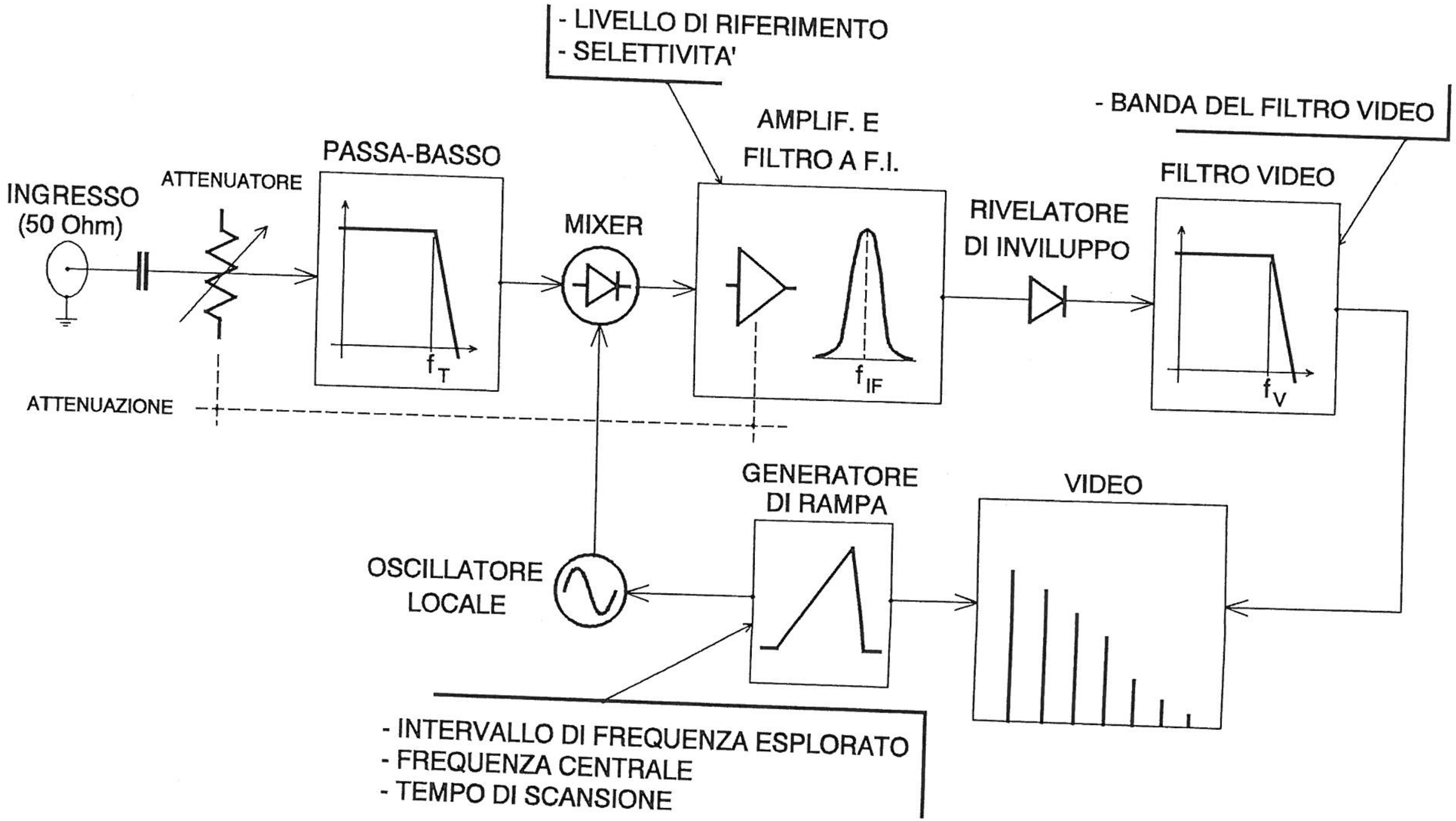
1 SWEEP
2 BW
3 TRIG
4 AUTO COUPLE
5 TRACE
6 DISPLAY

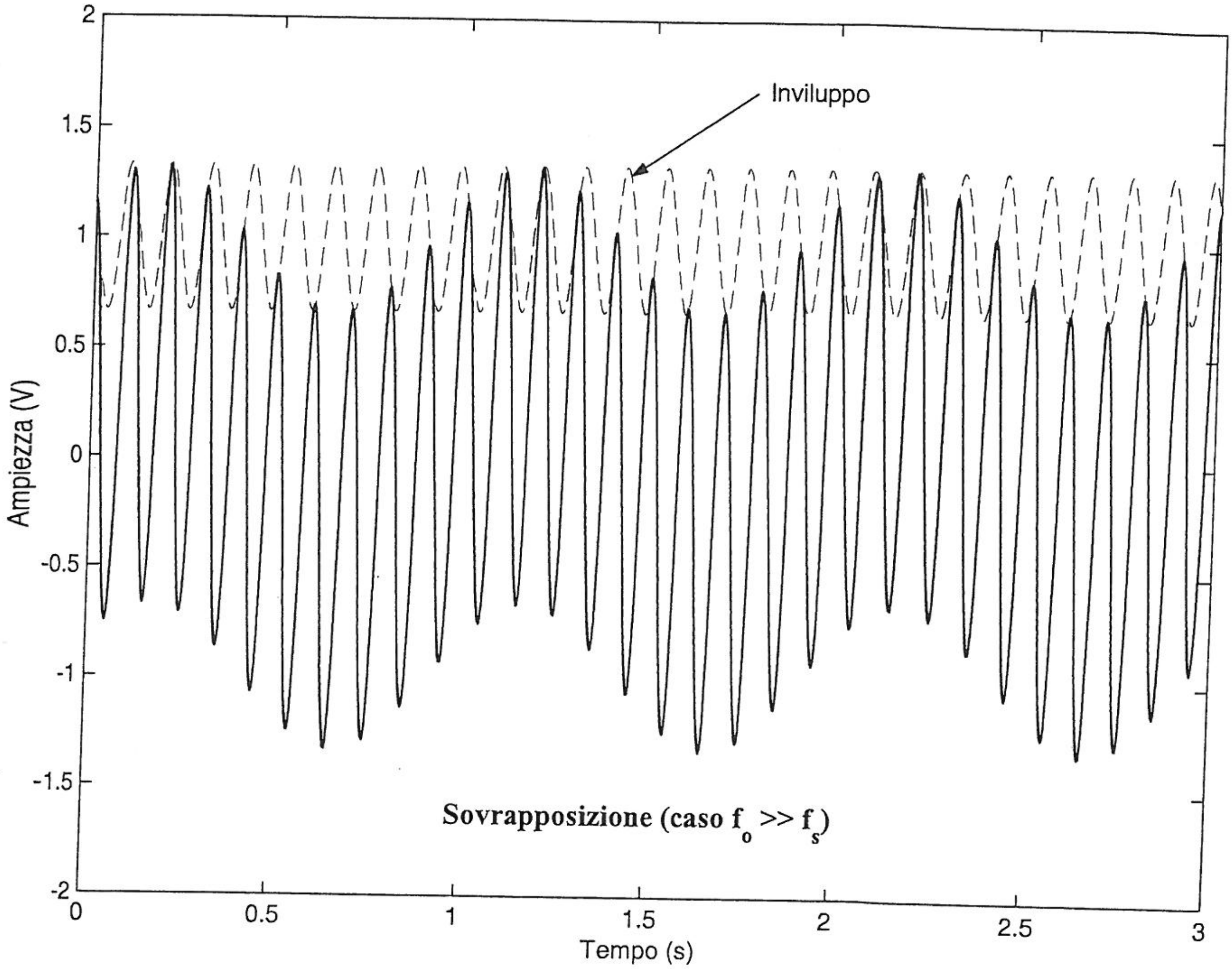
1 MKR
2 MKR→
3 MKR FCTN
4 PEAK SEARCH

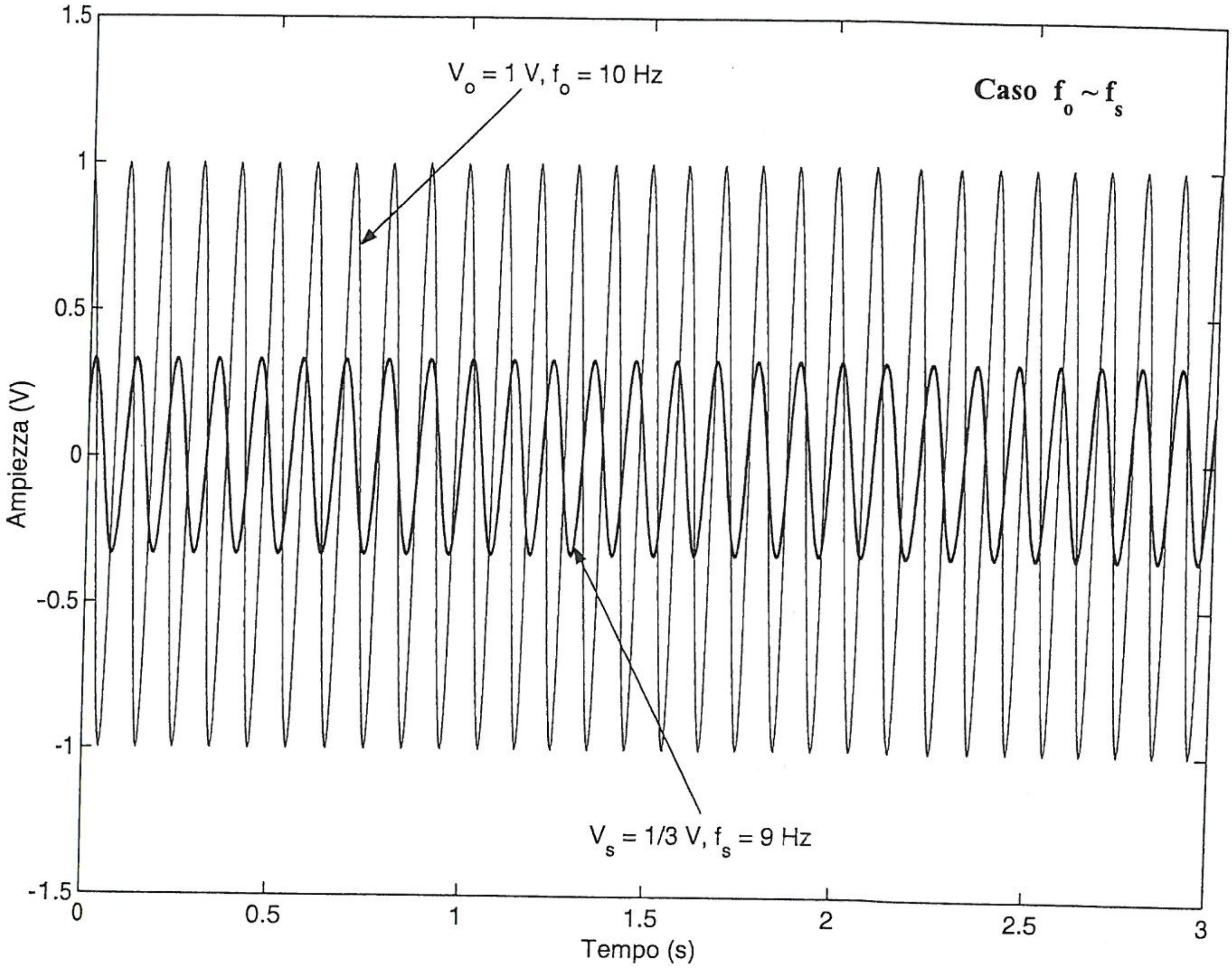
pu133c

Figure 2-1. Front-Panel Feature Overview

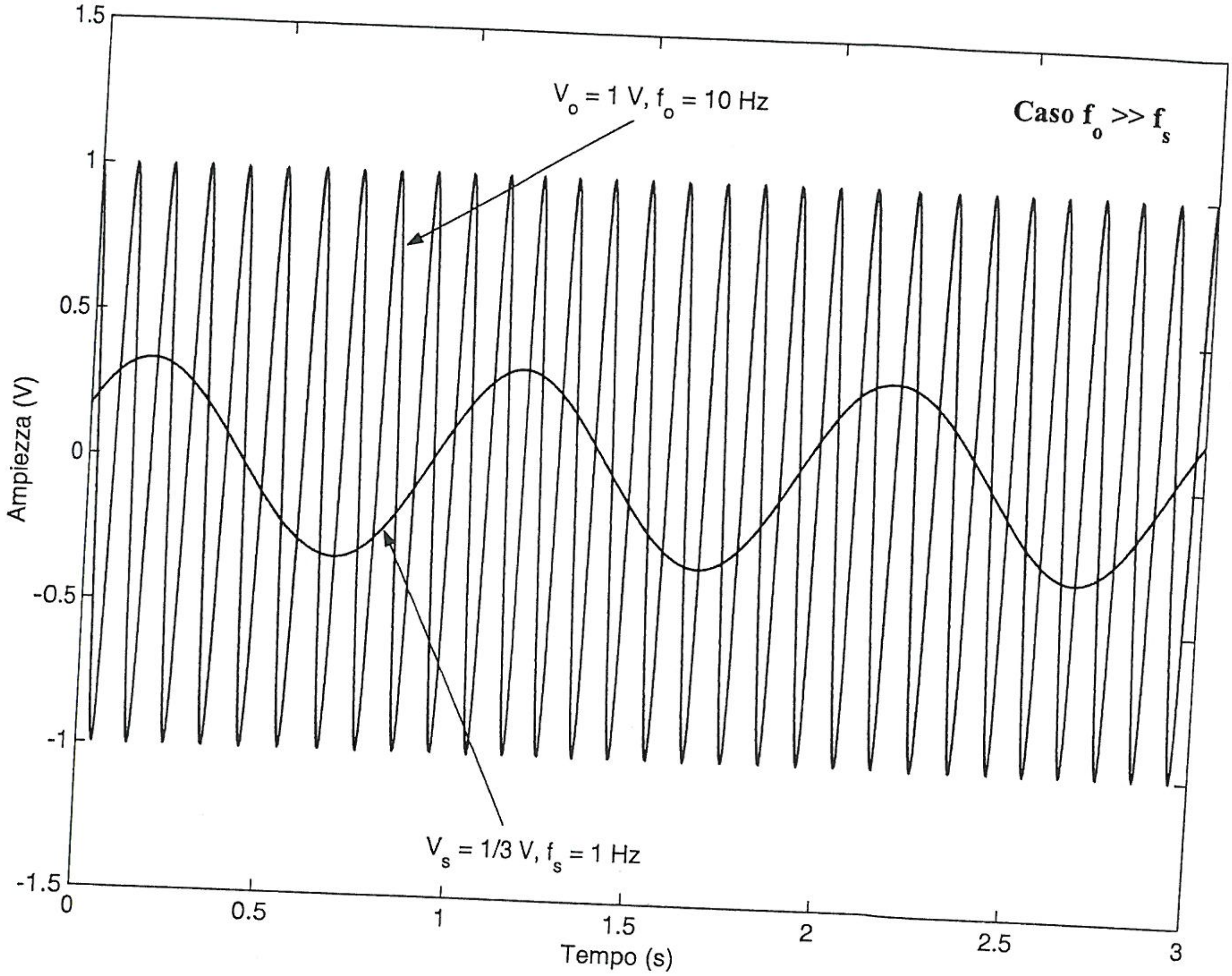
ANALIZZATORE DI SPETTRO - SCHEMA A BLOCCHI E COMANDI FONDAMENTALI



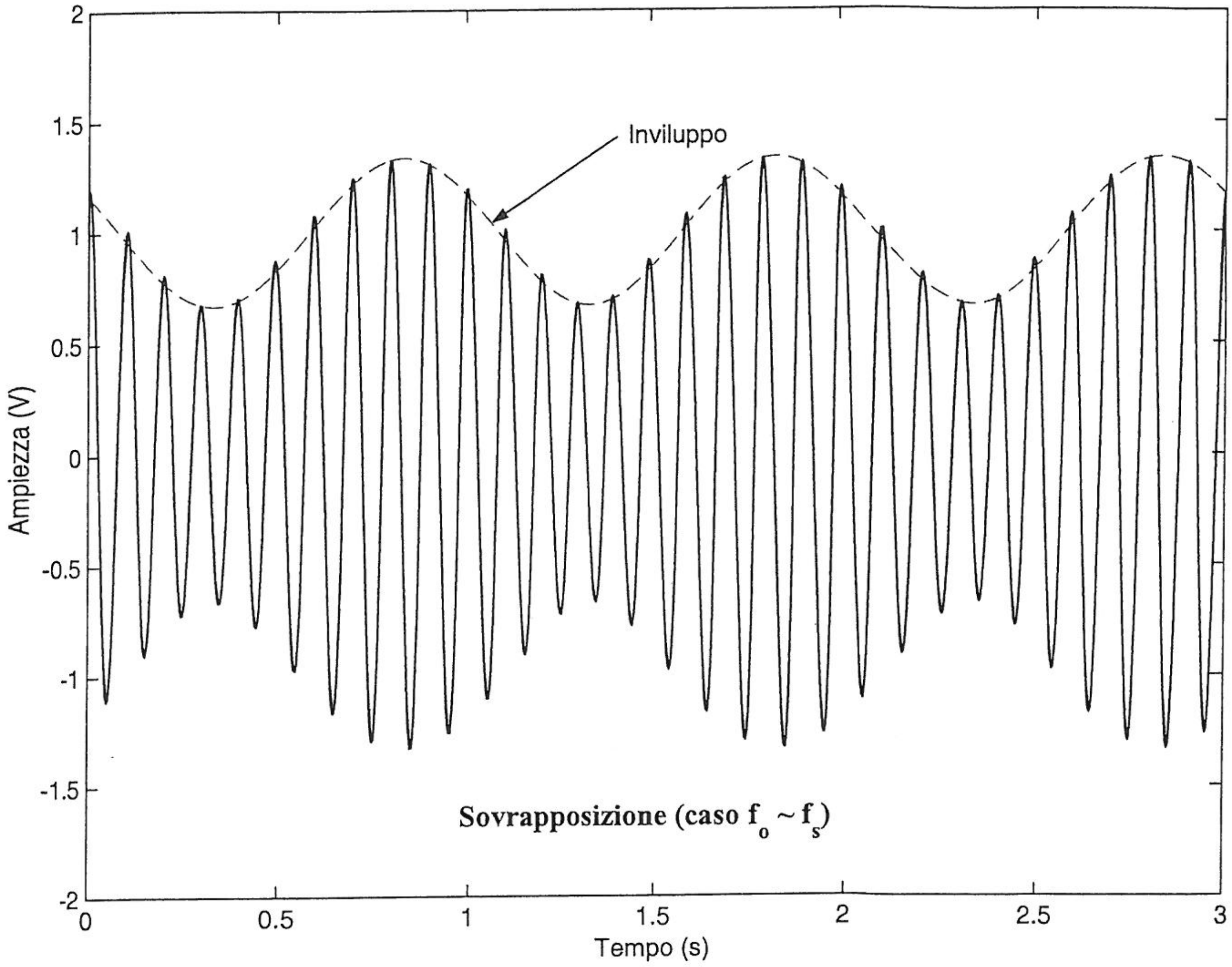

Inviluppo
Sovrapposizione (caso $f_o >> f_s$)
Ampiezza (V)
Tempo (s)
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5
3



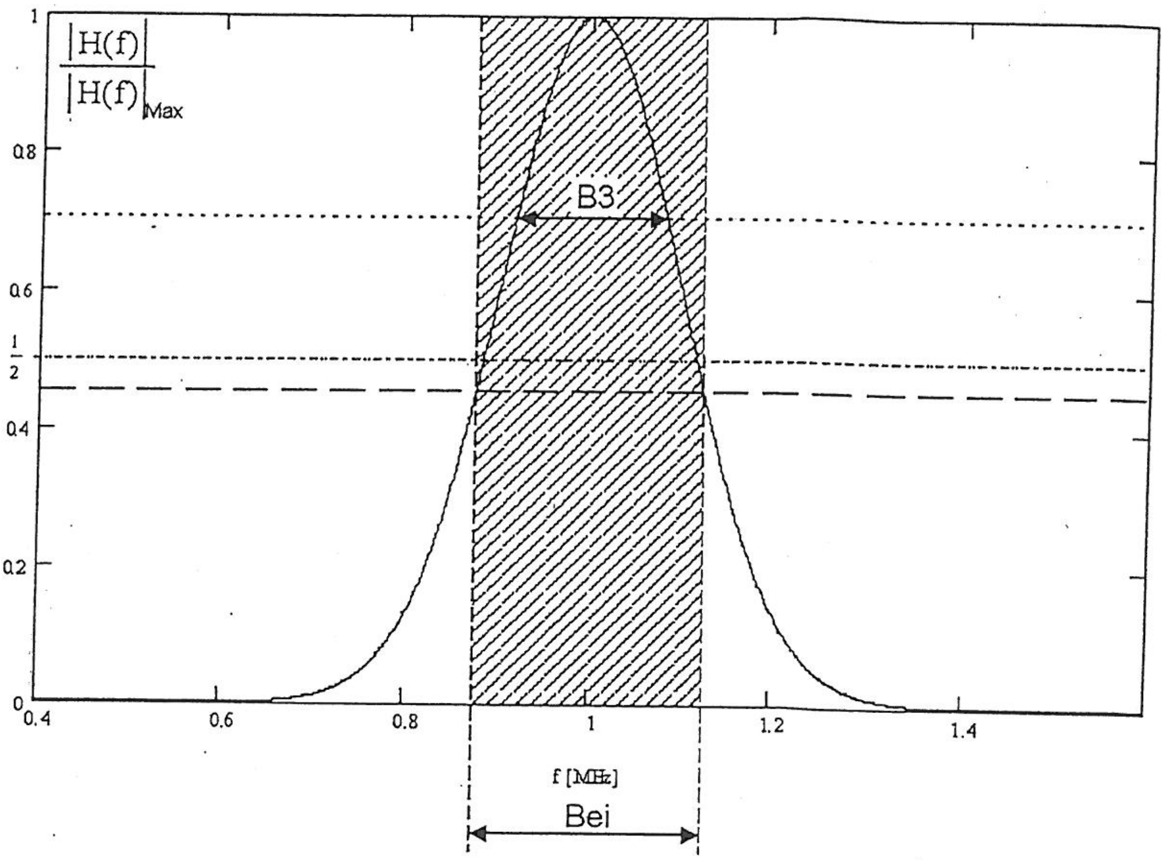
Caso $f_o \sim f_s$
V_o = 1 V, f_o = 10 Hz
V_s = 1/3 V, f_s = 9 Hz
1.5
1
0.5
0
-0.5
-1
-1.5
Ampiezza (V)
0
0.5
1
1.5
2
2.5
3
Tempo (s)



Caso $f_o >> f_s$
V_o = 1 V, f_o = 10 Hz
V_s = 1/3 V, f_s = 1 Hz
Ampiezza (V)
Tempo (s)
1.5
1
0.5
0
-0.5
-1
-1.5
0
0.5
1
1.5
2
2.5
3



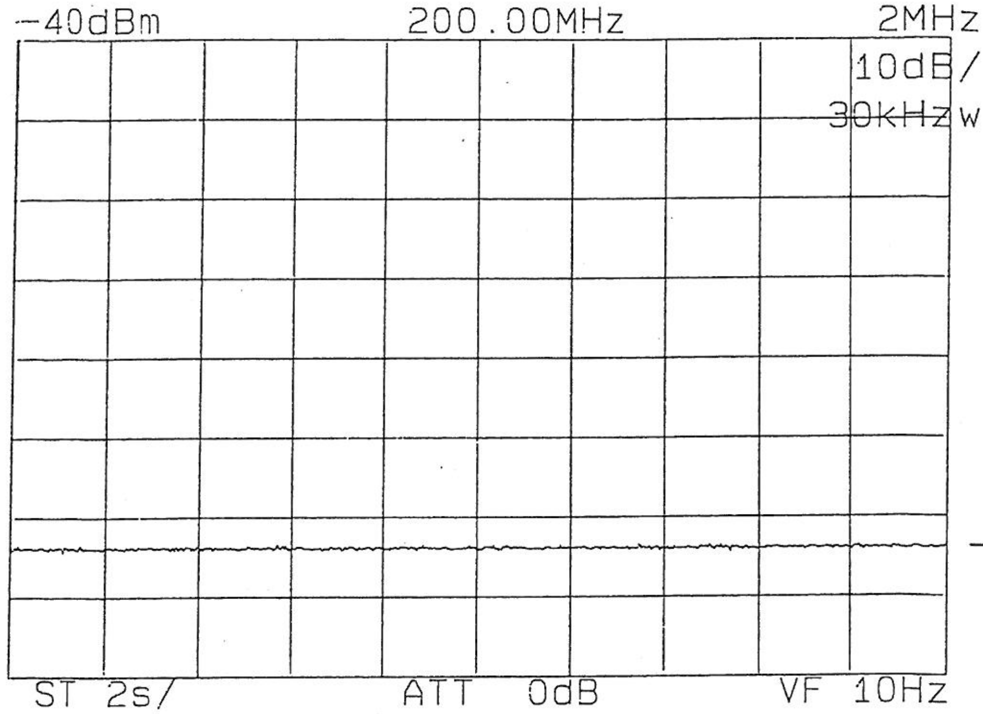
Inviluppo
Sovrapposizione (caso $f_o \sim f_s$)
Ampiezza (V)
Tempo (s)
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
0
0.5
1
1.5
2
2.5
3



RISPOSTA IN FREQUENZA GAUSSIANA

LIVELLO DI RUMORE MEDIO SPETTROANALIZZATORE

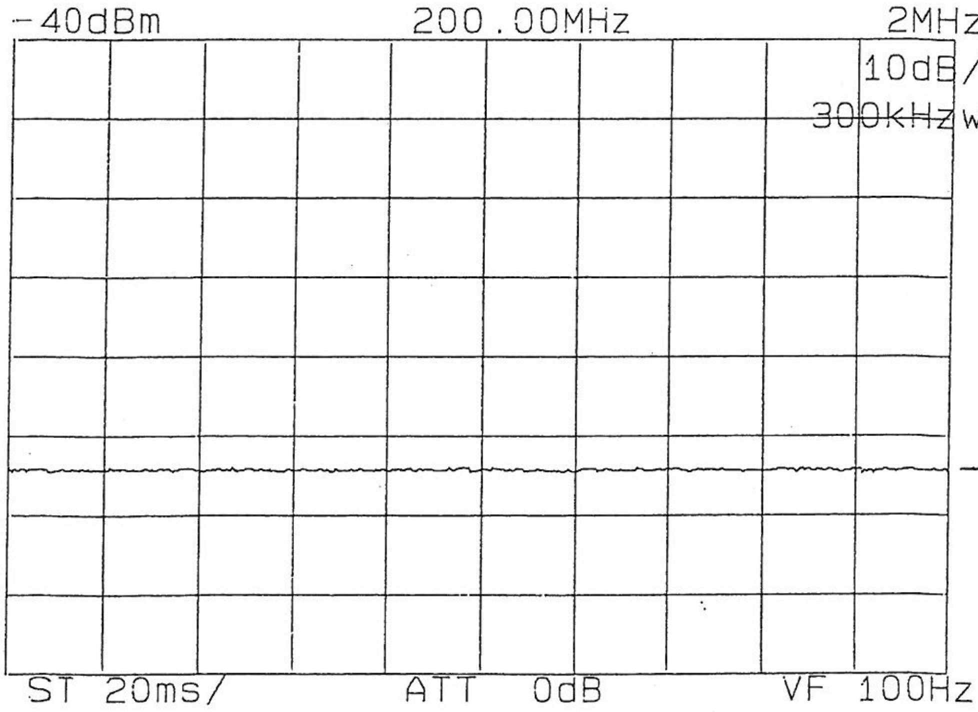
EFFETTO DELLA VARIAZIONE DI B_3



$B_{er} \simeq 35$ kHz

$F \simeq 27.6$ dB

— -103.4 dBm



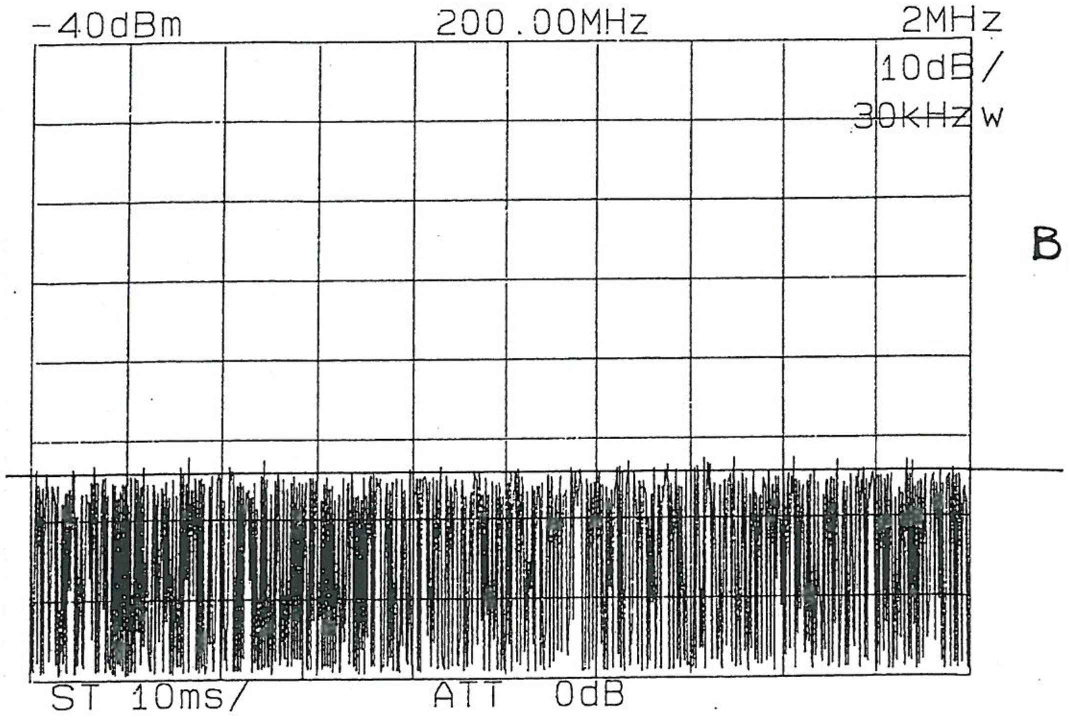
$B_{er} \simeq 356$ kHz

$F \simeq 27.0$ dB

— -94.0 dBm

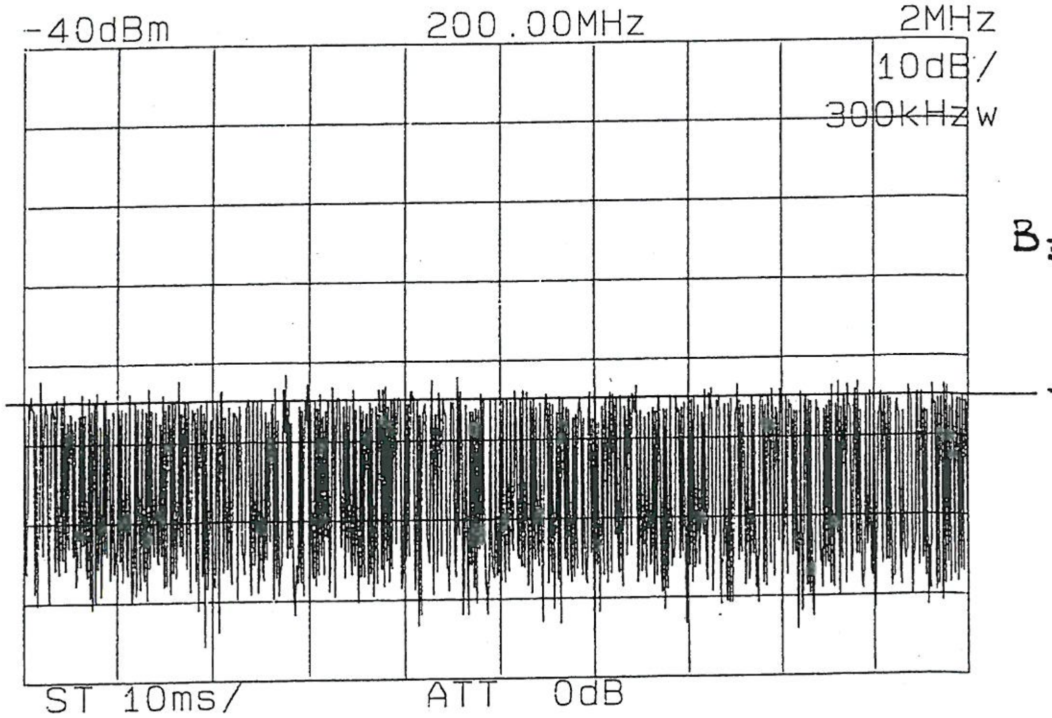
DISTURBO A LARGA BANDA INCOERENTE (RUMORE TERMICO)

EFFETTO DELLA VARIAZIONE DI B_3



$B_3 = 30$ kHz (NOM.)

-94.4 dBm

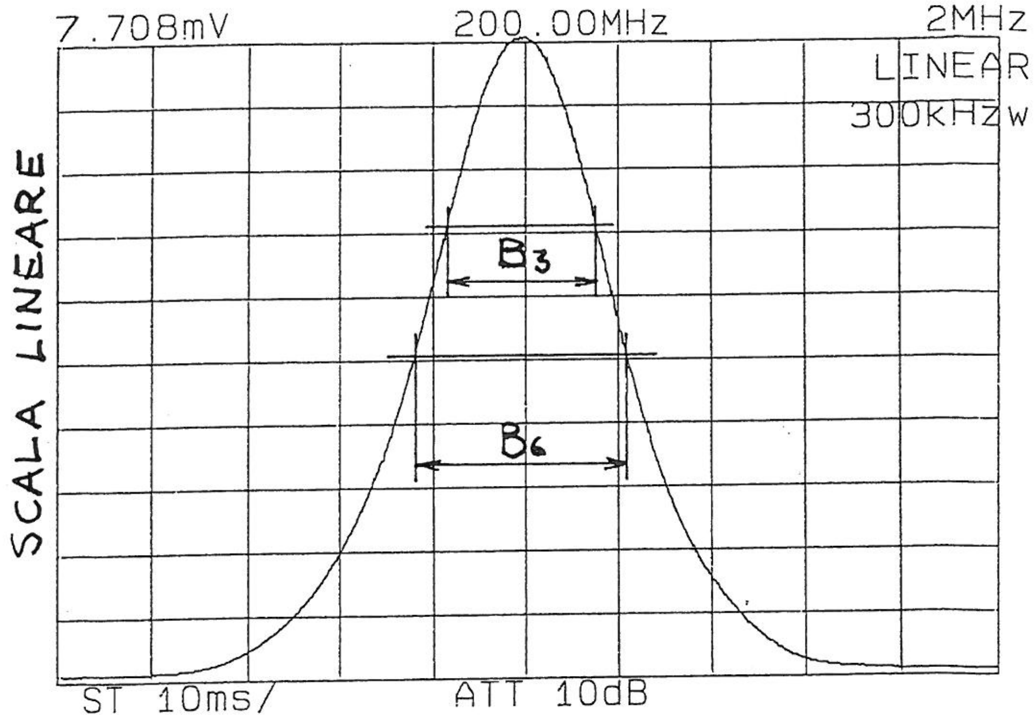


$B_3 = 300$ kHz (NOM.

-84.5 dBm

RISPOSTA IN FREQUENZA FILTRO IF DELLO SPETTRO ANALIZZATORE

FILTRO $B_3 = 300$ kHz (NOM.)

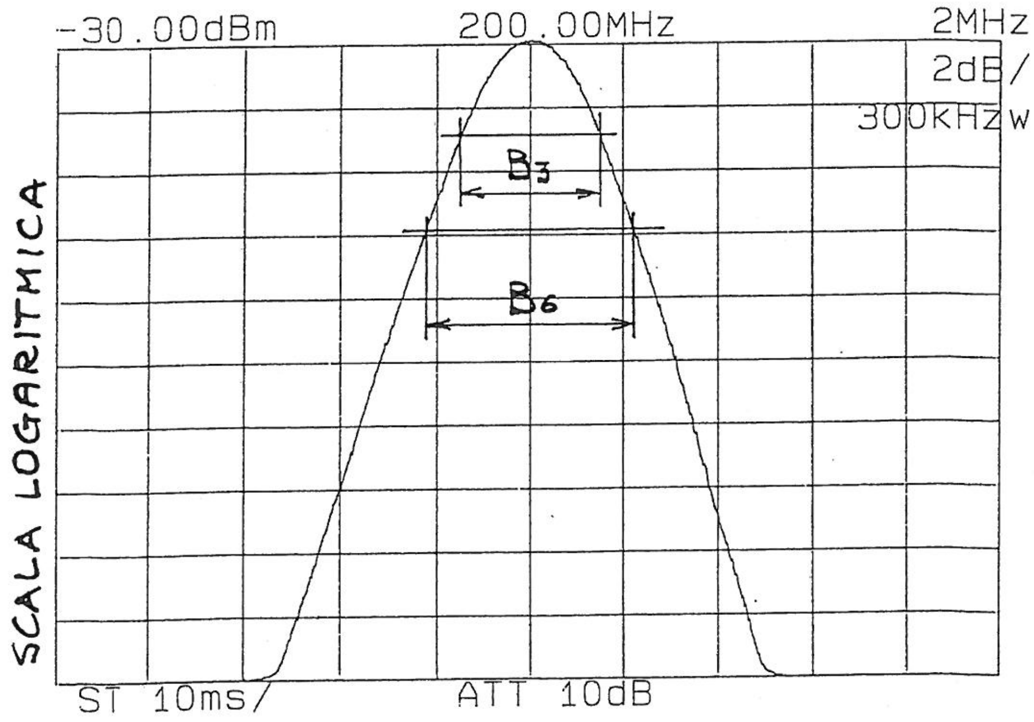


$B_3 \simeq 324$ kHz

$B_6 \simeq 469$ kHz

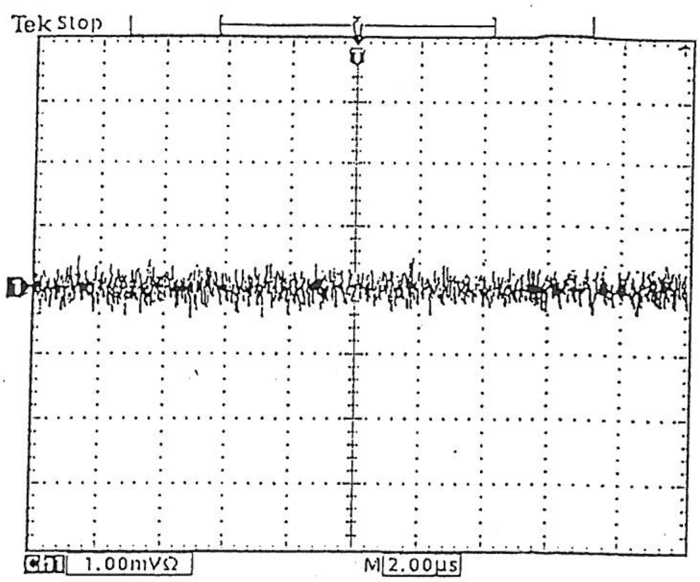
$B_{ei} \simeq 524$ kHz

$B_{en} \simeq 356$ kHz



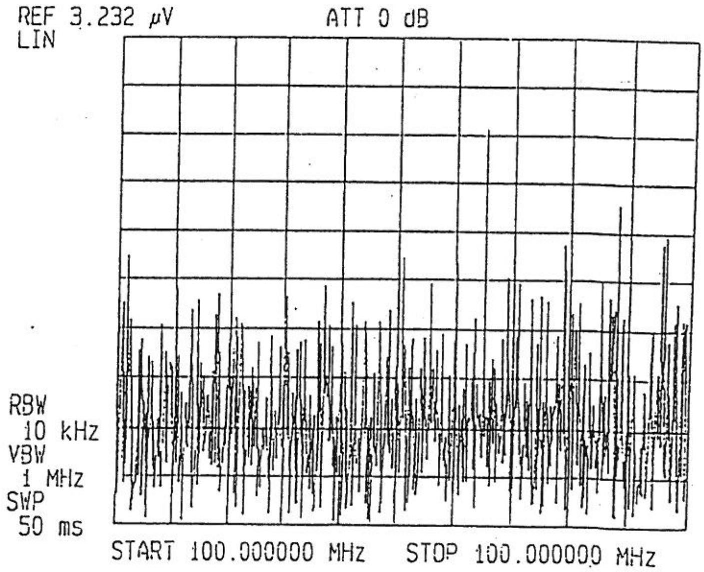
RUMORE – PRESENTAZIONE OSCILLOSCOPICA

OSCILLOSCOPIO

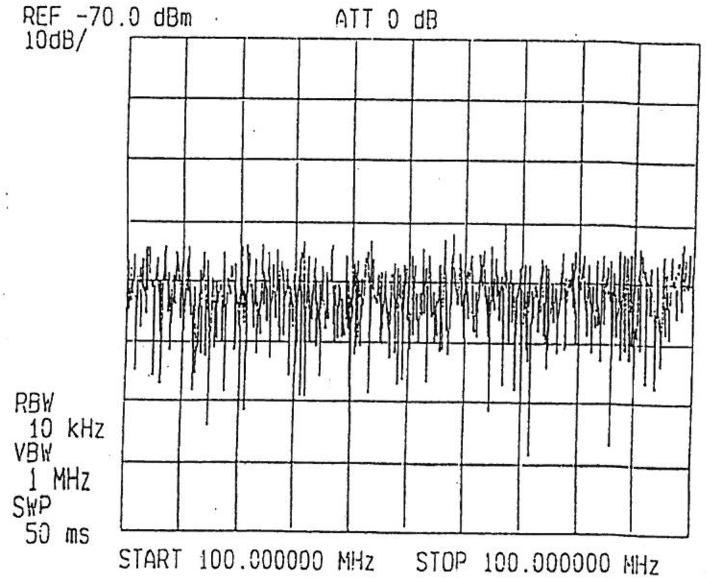


SPETTROANALIZZATORE

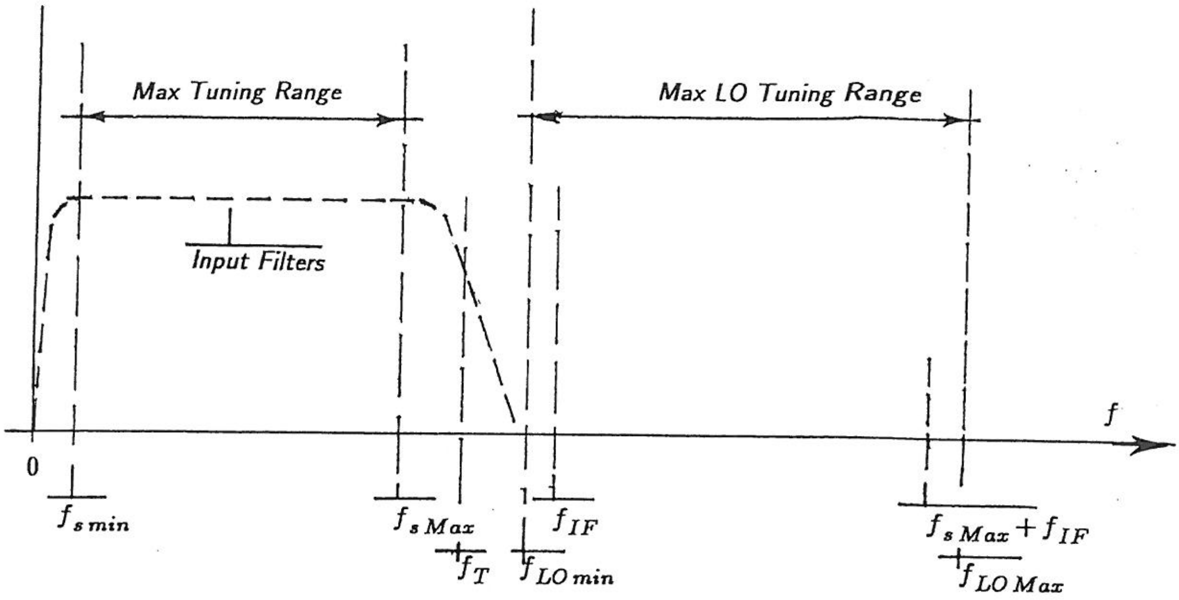
SCALA VERTICALE LINEARE



SCALA VERTICALE LOGARITMICA



SPETTROANALIZZATORE: MESCOLAZIONE IN FONDAMENTALE



Si impone che la frequenza intermedia sia al di fuori della gamma di segnale che si intende osservare. In questo caso si sceglie:

$$f_{s\,Max} < f_{IF}$$

condizione garantita dal filtro passa-basso, taglio a f_T:

$$f_{s\,Max} < f_T < f_{IF}$$

Vogliamo anche:

$$f_{s\,min} \simeq 0$$

Per coprire l'intervallo $f_{s\,min} \leq f_s \leq f_{s\,Max}$ deve aversi:

$$f_{s\,min} + f_{IF} = f_{LO\,min} \leq f_{LO} \leq f_{LO\,Max} \geq f_{s\,Max} + f_{IF}$$

In pratica si adotta:

$$f_{LO\,min} < \simeq f_{s\,min} + f_{IF}$$

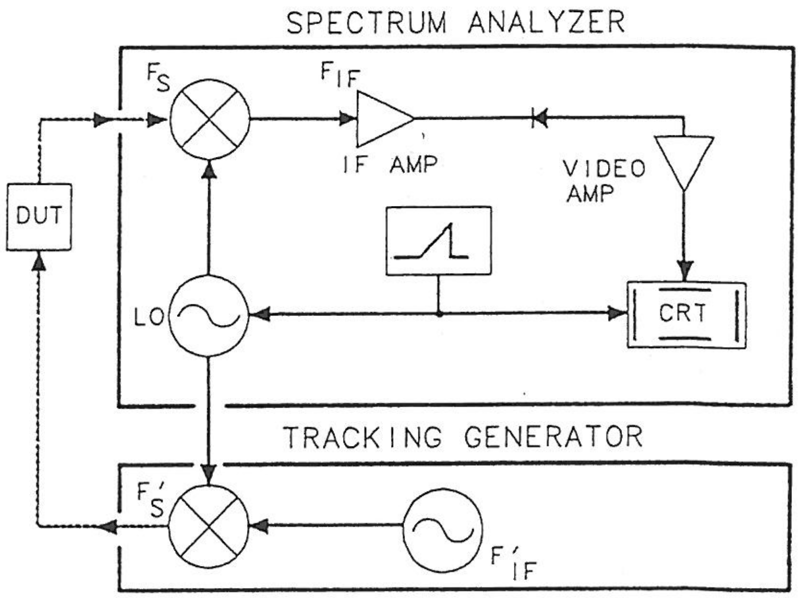
(per cui il segnale del *LO* appare sul visore all'estremo sinistro della gamma di sintonia)

Infine: $f_{s\,min} > \simeq 0 \ll f_{IF}$ (con filtro passa-alto, $f_{taglio} < f_{s\,min}$)

Generatore Tracking

Consente di effettuare misure scalari di perdita di inserzione (*insertion loss*), funzione di trasferimento, coefficiente di riflessione (insieme a un ponte a radiofrequenza).

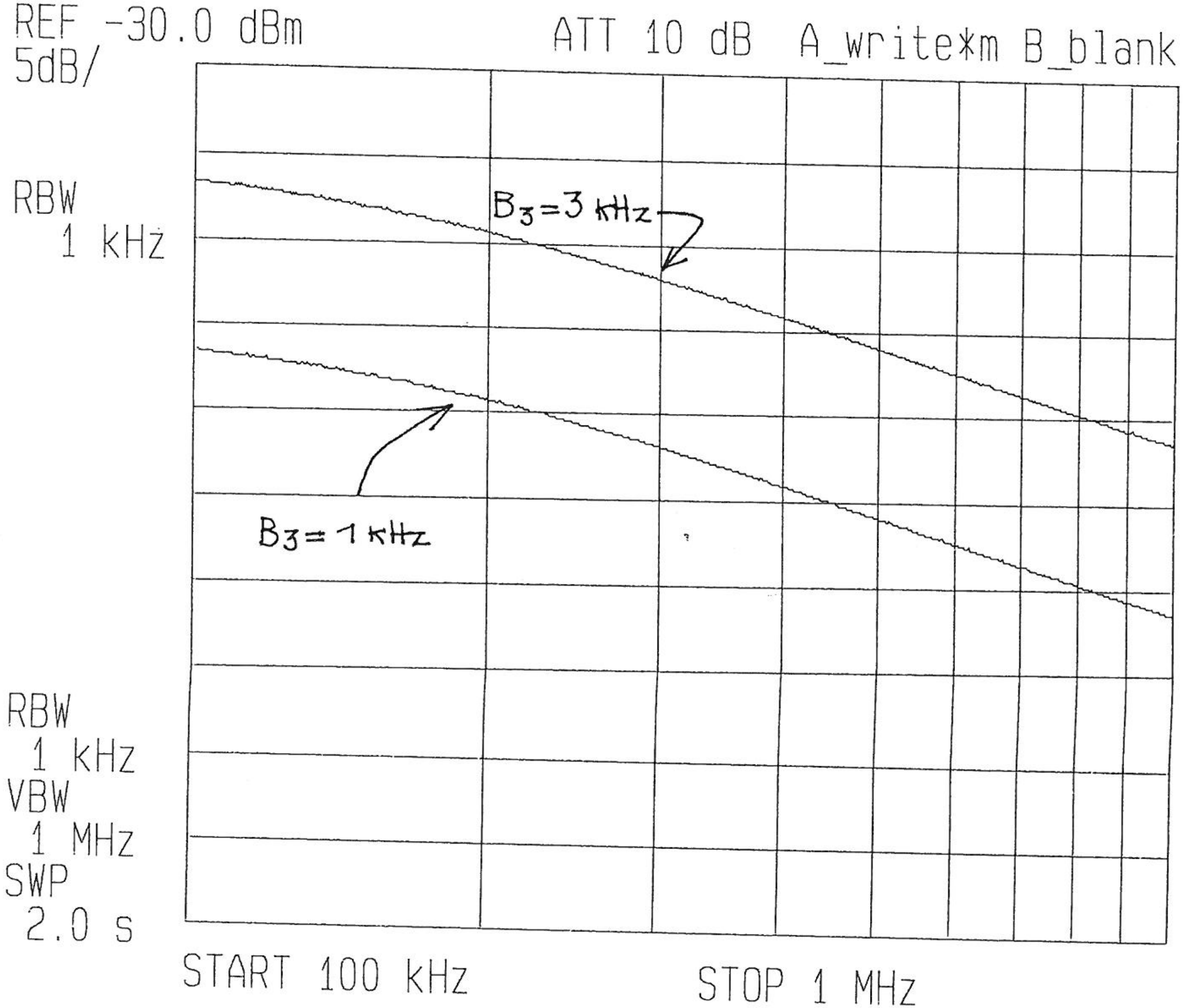
Principio di funzionamento (DUT = Device Under Test):



$$\boxed{F'_S = F_{LO} - F'_{IF} = F_{LO} - F_{IF}}$$

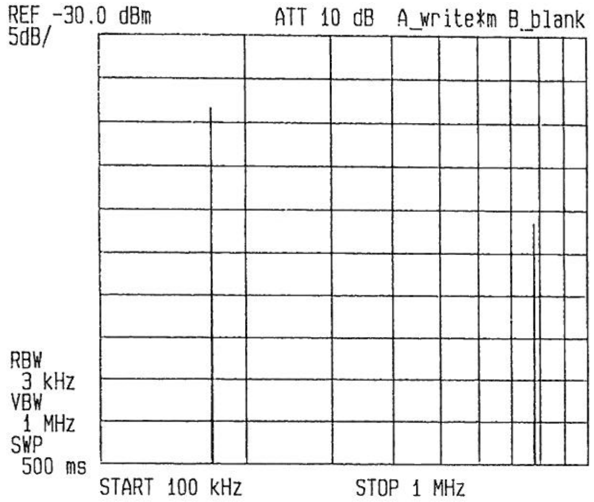
DISTURBO A LARGA BANDA COERENTE (SCARICA CAPACITA')

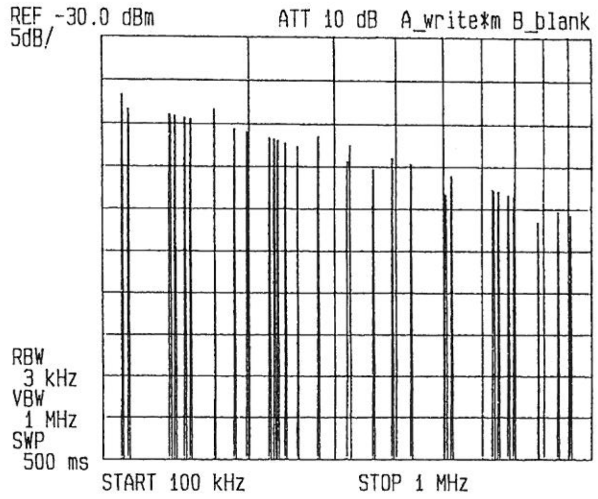
EFFETTO DELLA VARIAZIONE DELLA B_3

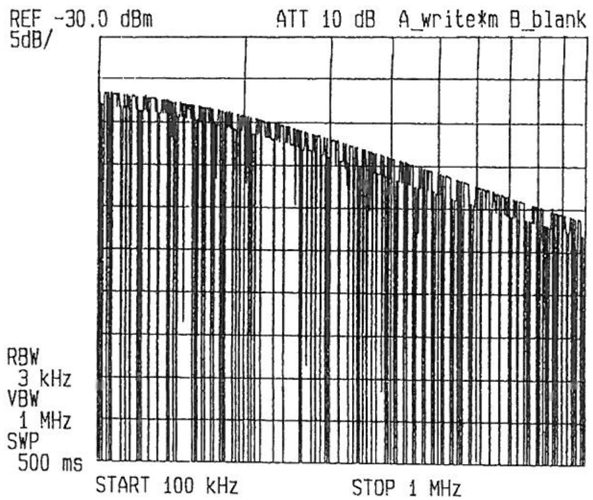


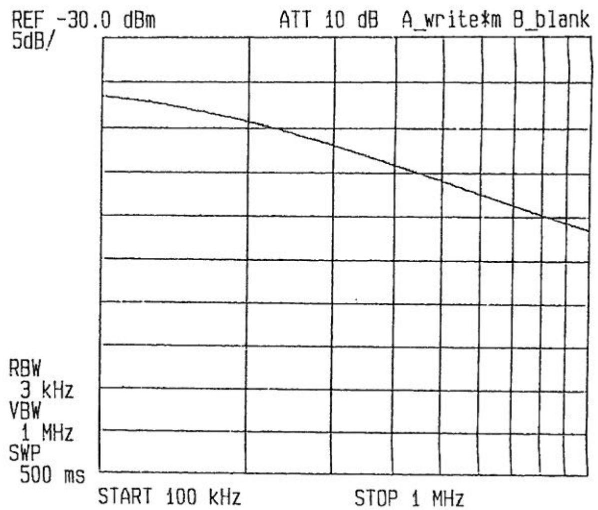
SPETTRO DELLA TENSIONE DI SCARICA DI UNA CAPACITÀ

$v(t) = A e^{-t/\tau}$; $\tau = RC$ $(A = 2/\sqrt{10}\ V,\ R = 100\Omega,\ C = 10\,nF)$





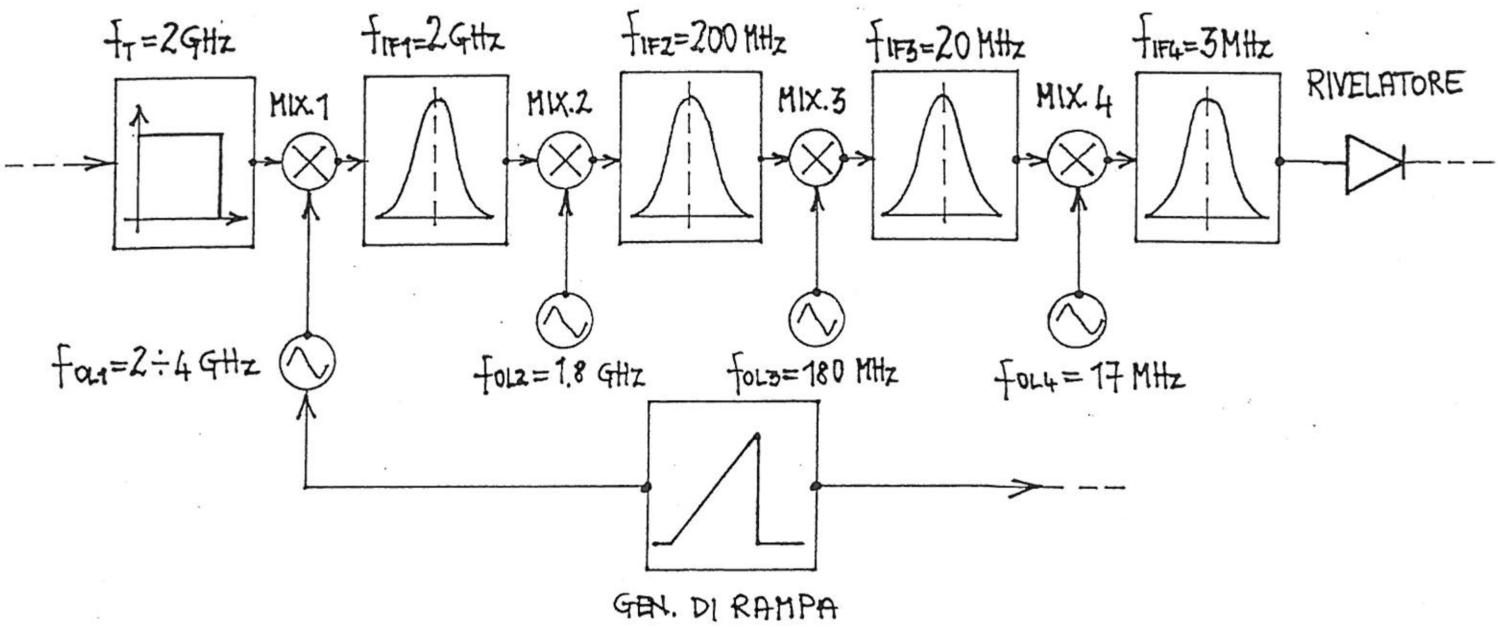




Panoramicità e selettività

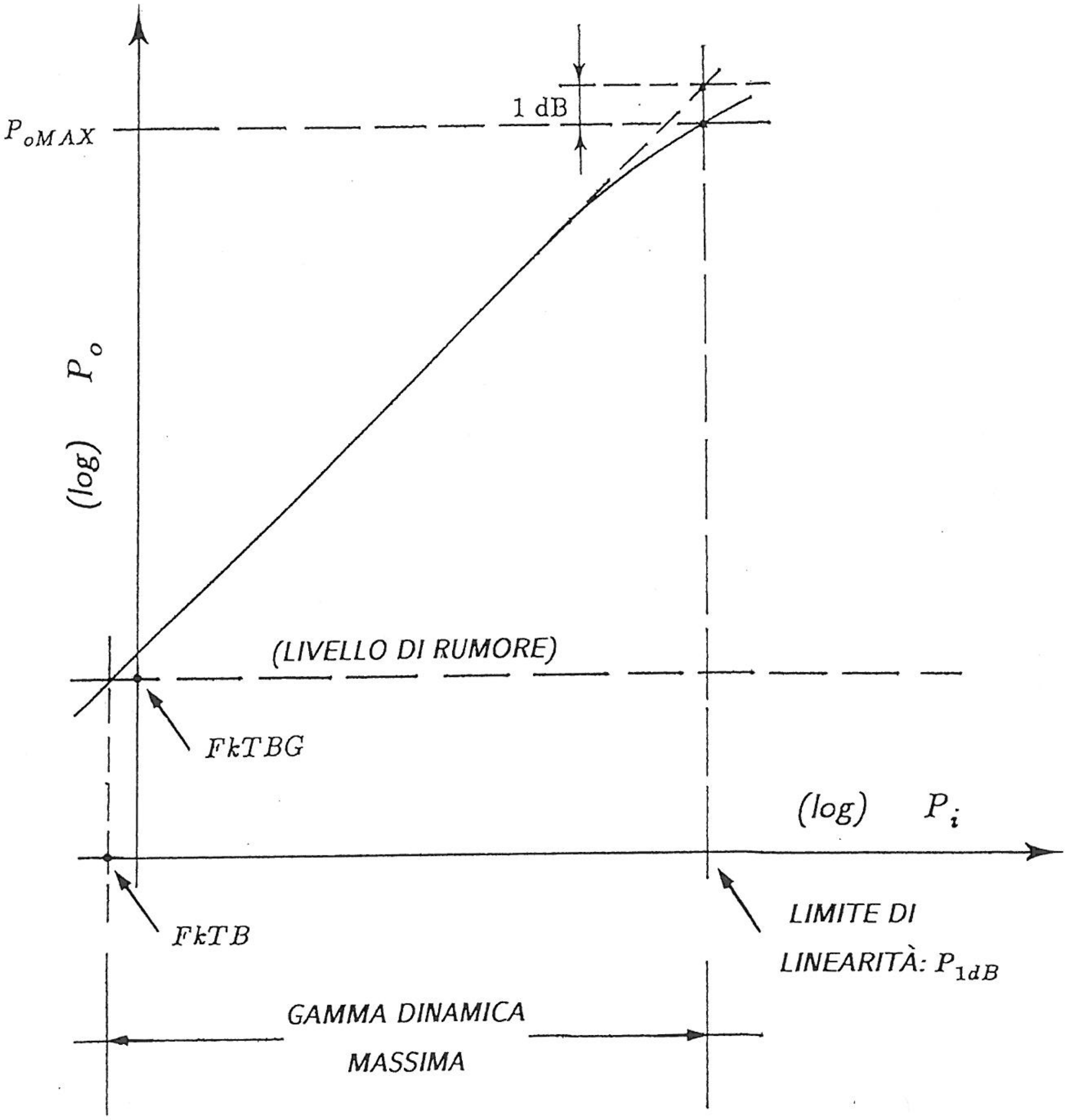
Per ottenere *panoramicità* (alta $f_{IF} \simeq f_{S\,MAX}$) e *selettività* (B_3 stretta) sono necessari *più stadi di conversione.*

Esempio:



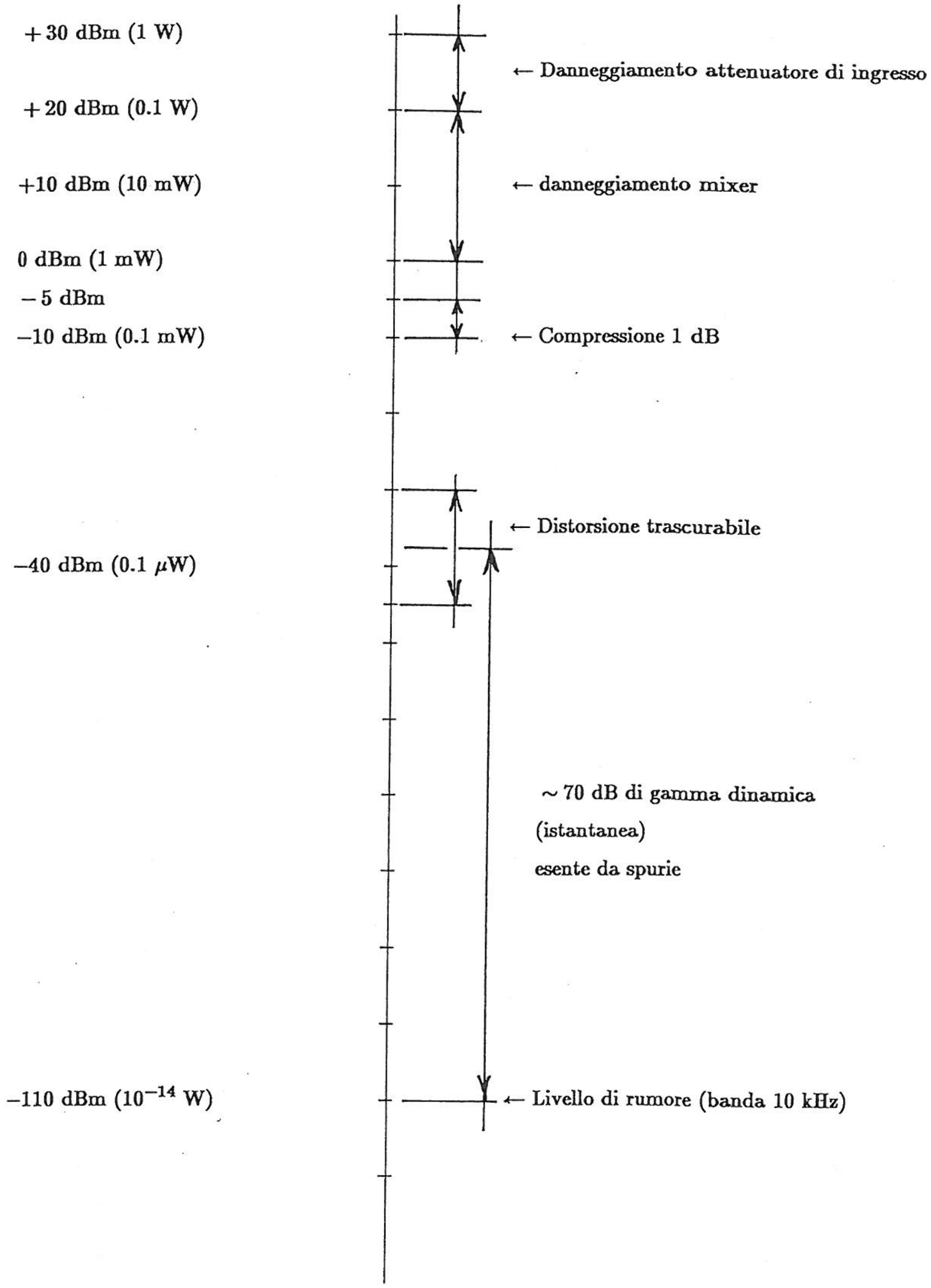
▷ L'ultimo filtro a frequenza intermedia ha banda B_3 variabile da pannello frontale (sequenza $1 - 3 - 10$) e determina la selettività dello strumento.

▷ La prima frequenza intermedia determina la panoramicità (intervallo di frequenze di sintonia)



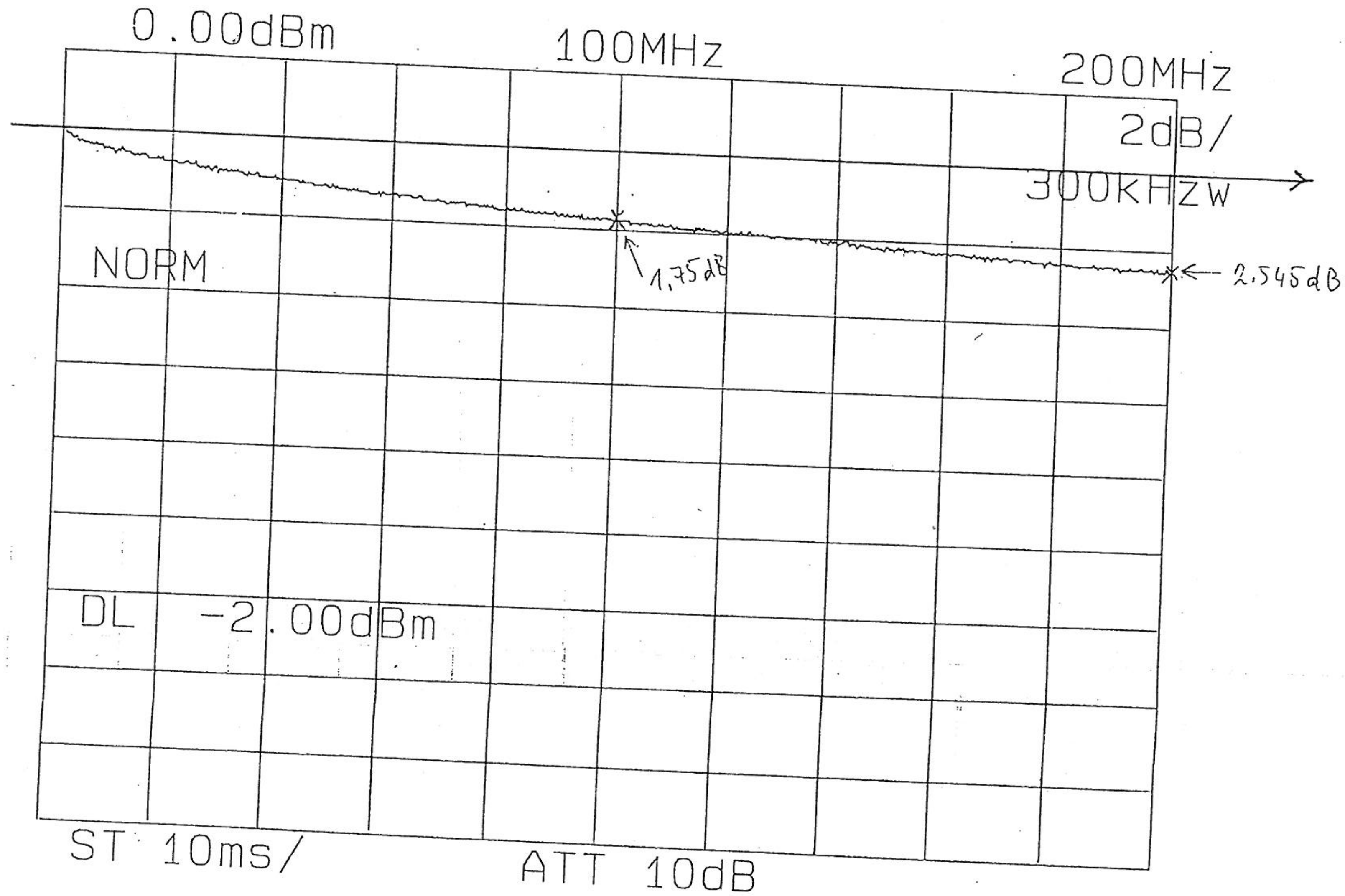
P_{oMAX}
1 dB
(log) P_o
(LIVELLO DI RUMORE)
$FkTBG$
(log) P_i
$FkTB$
LIMITE DI
LINEARITÀ: P_{1dB}
GAMMA DINAMICA
MASSIMA

LIVELLI TIPICI DI SEGNALE NELLO SPETTROANALIZZATORE

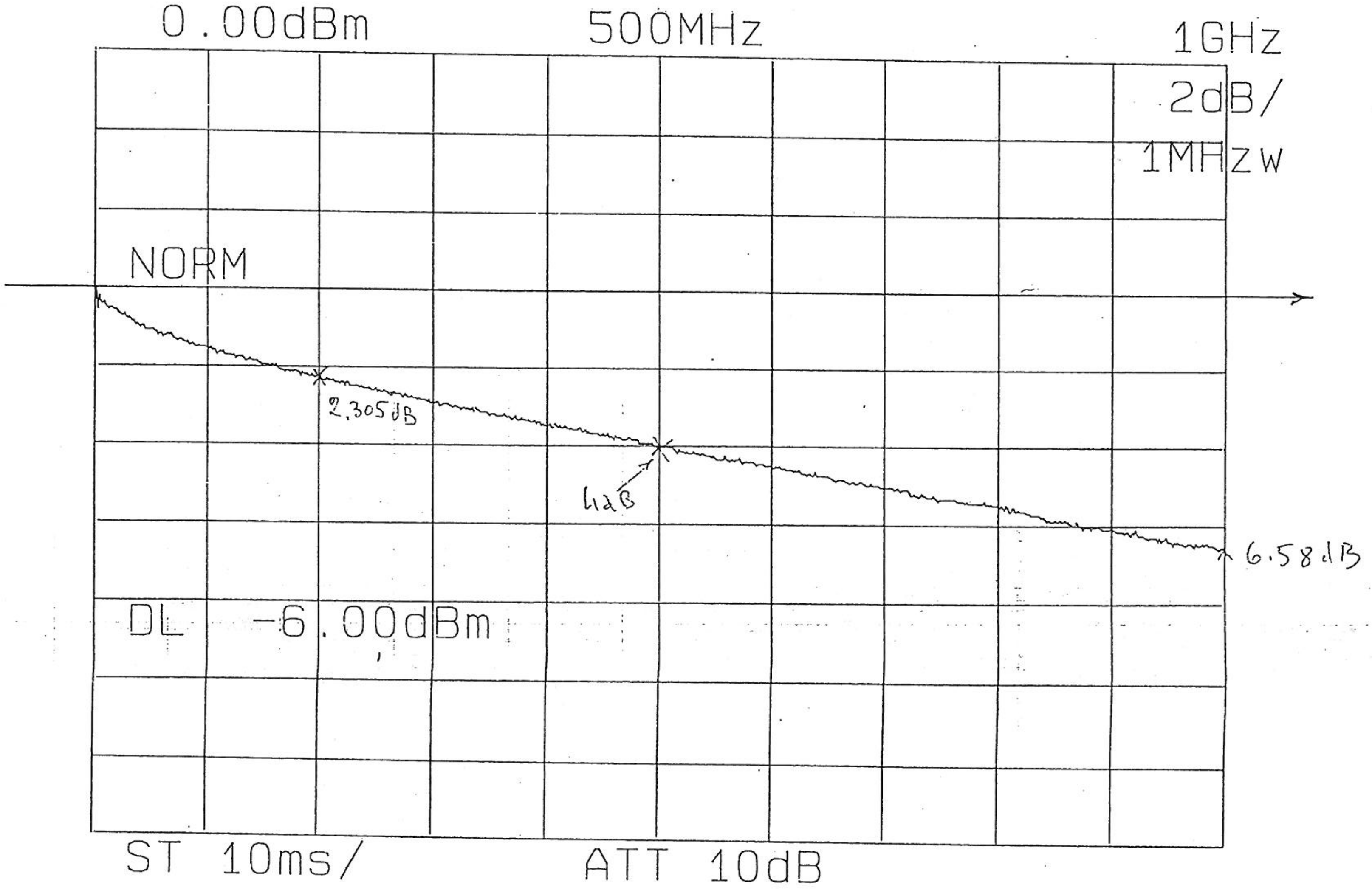


10 m

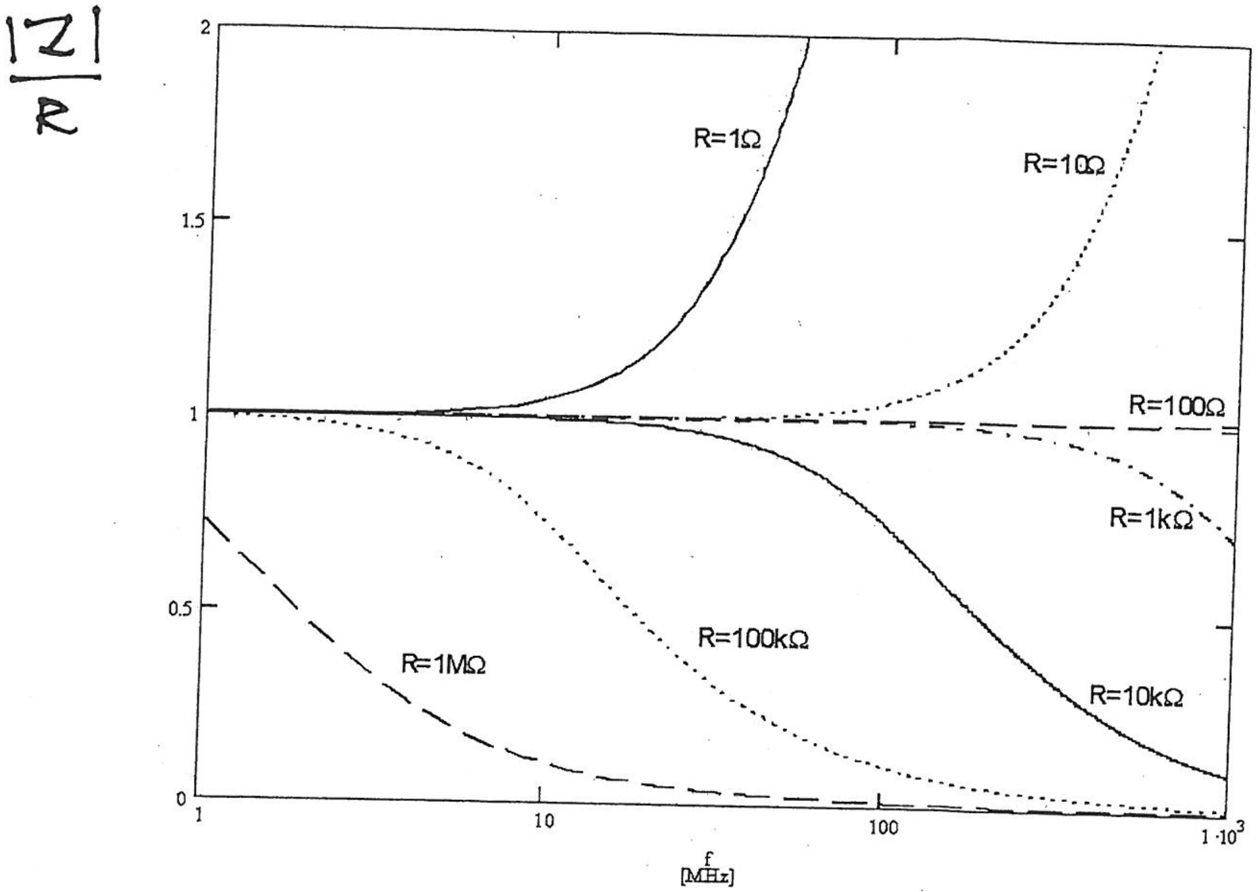
12 OTT. '98


0.00dBm
100MHz
200MHz
2dB/
300KHZW
NORM
1,75dB
2,545dB
DL -2.00dBm
ST 10ms/
ATT 10dB

CAVO RG-58C/U 10 m 26 MAR. '98

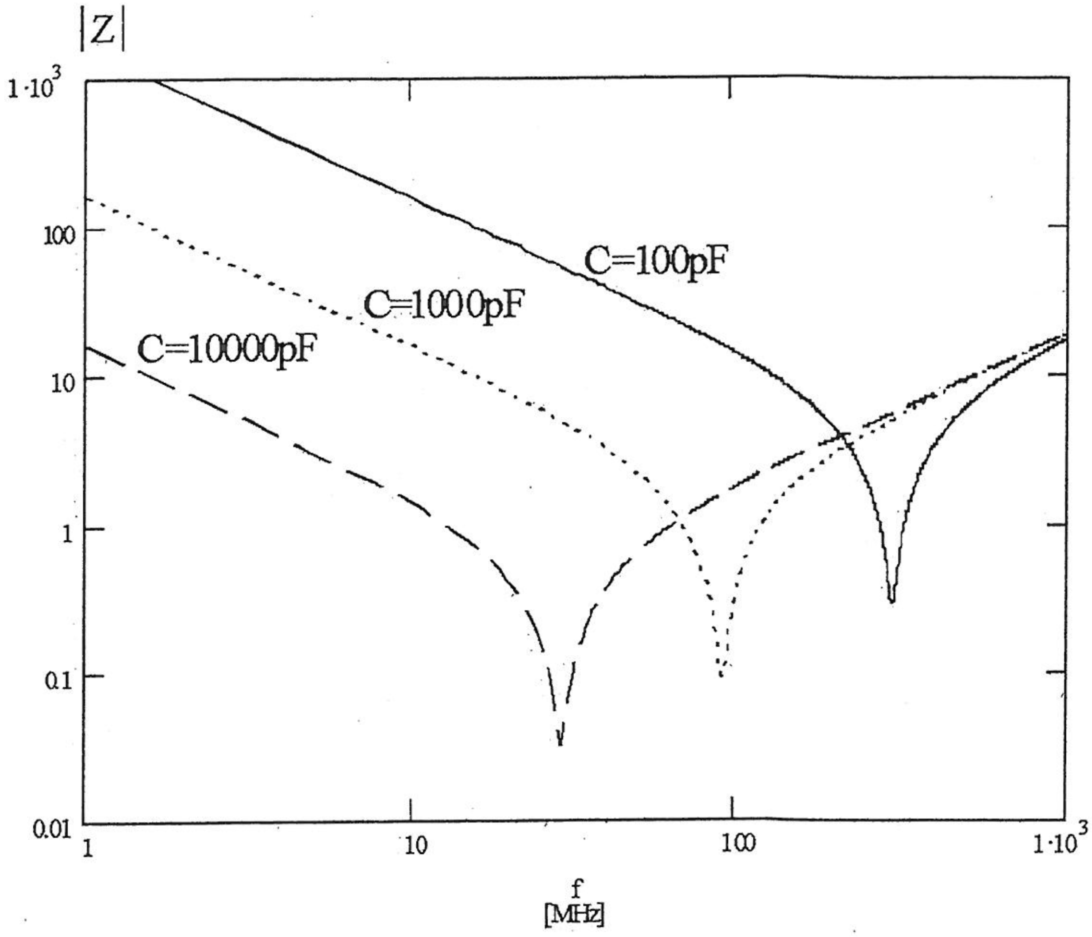


NON IDEALITA' DEL RESISTORE



$L = 5\text{nH}$
$C = 0.15\text{pF}$

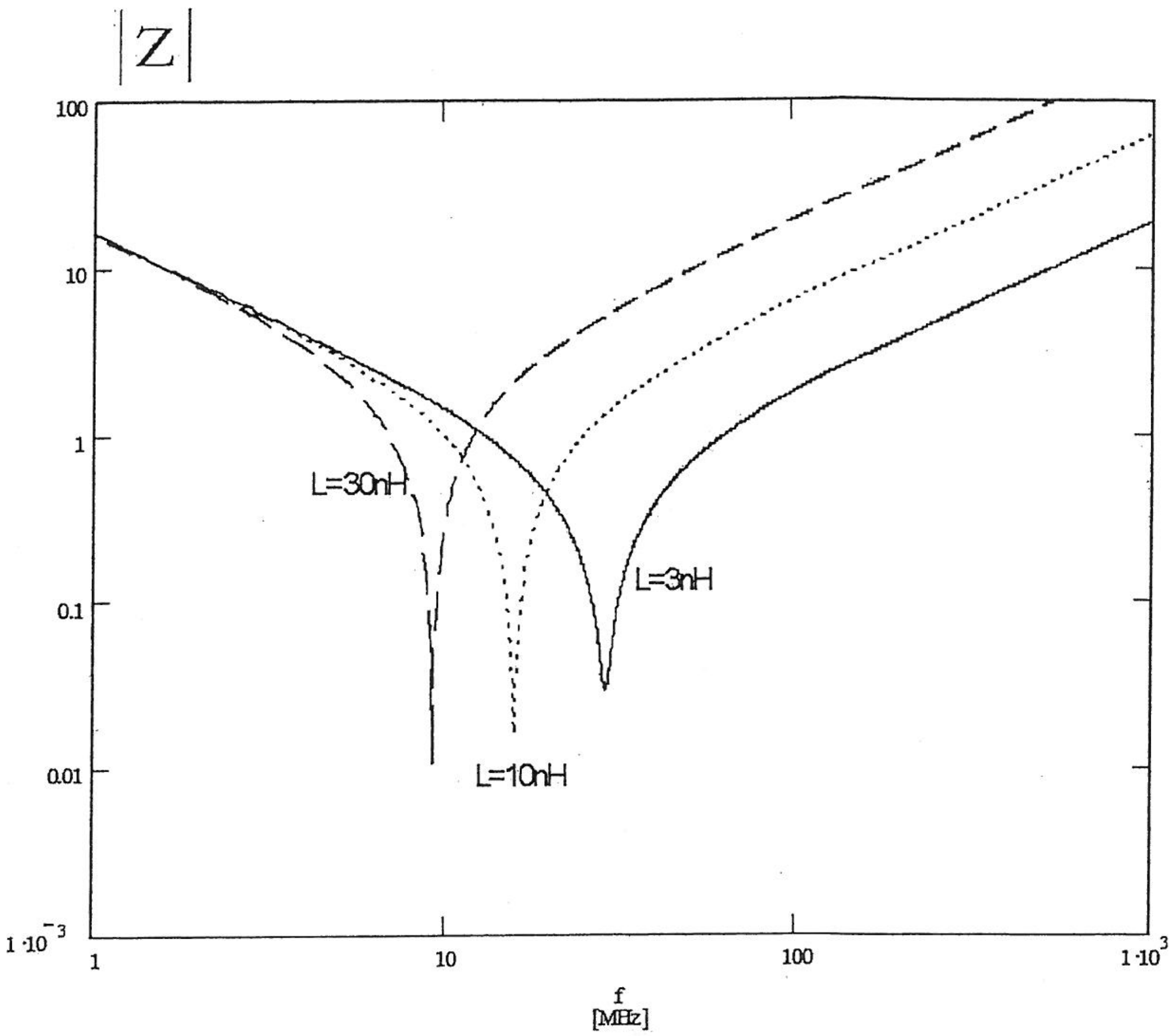
NON IDEALITA' DEL CONDENSATORE (1 DI 2)



$L = 3\ nH$

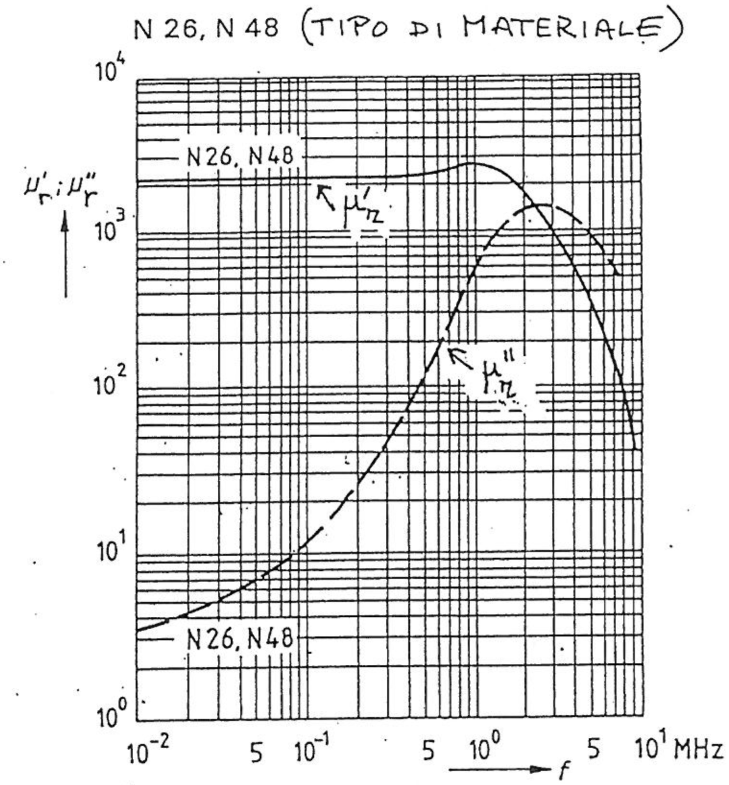
$R = 10^{-9} * f\ \ \Omega$ (f in Hz, R in Ω)

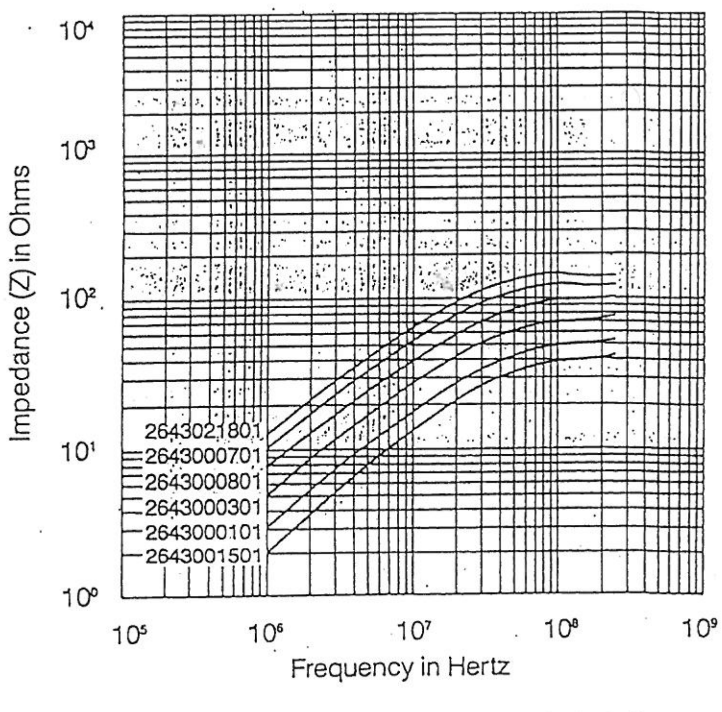
NON IDEALITÀ DEL CONDENSATORE (2 DI 2)



$C = 10000\text{pF}$ $R = 10^{-9} * f$ (f in Hz, R in Ω)

CARATTERISTICHE MAGNETICHE FERRITI
(ESEMPI DA CATALOGHI SIEMENS E FAIR-RITE)





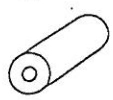
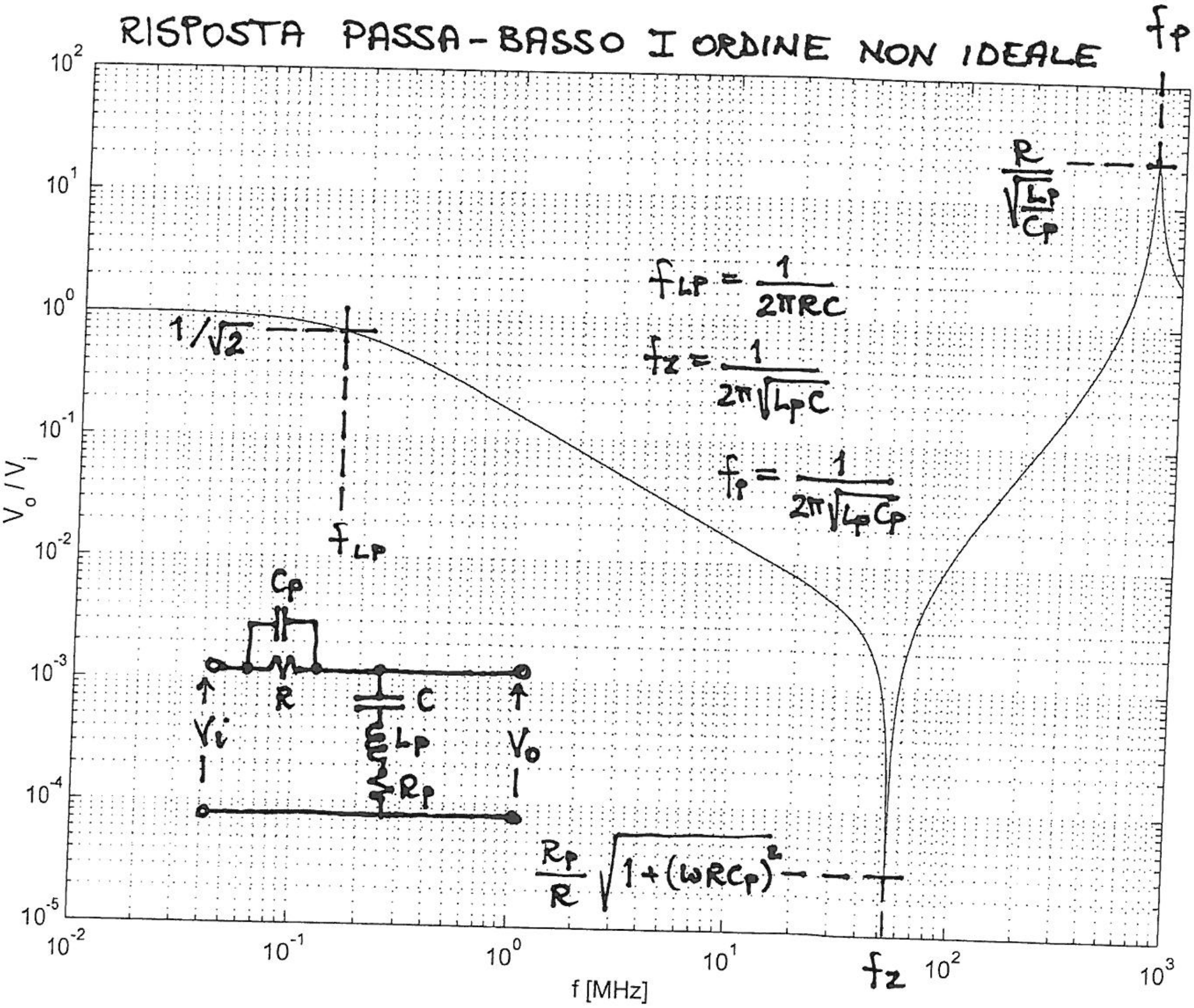


Figure 15. Impedance vs. Frequency of some single hole shield beads in Fair-Rite 43 material. 43 material is a nickel zinc material with medium permeability and high resistivity, recommended for frequencies up to 200 MHz.



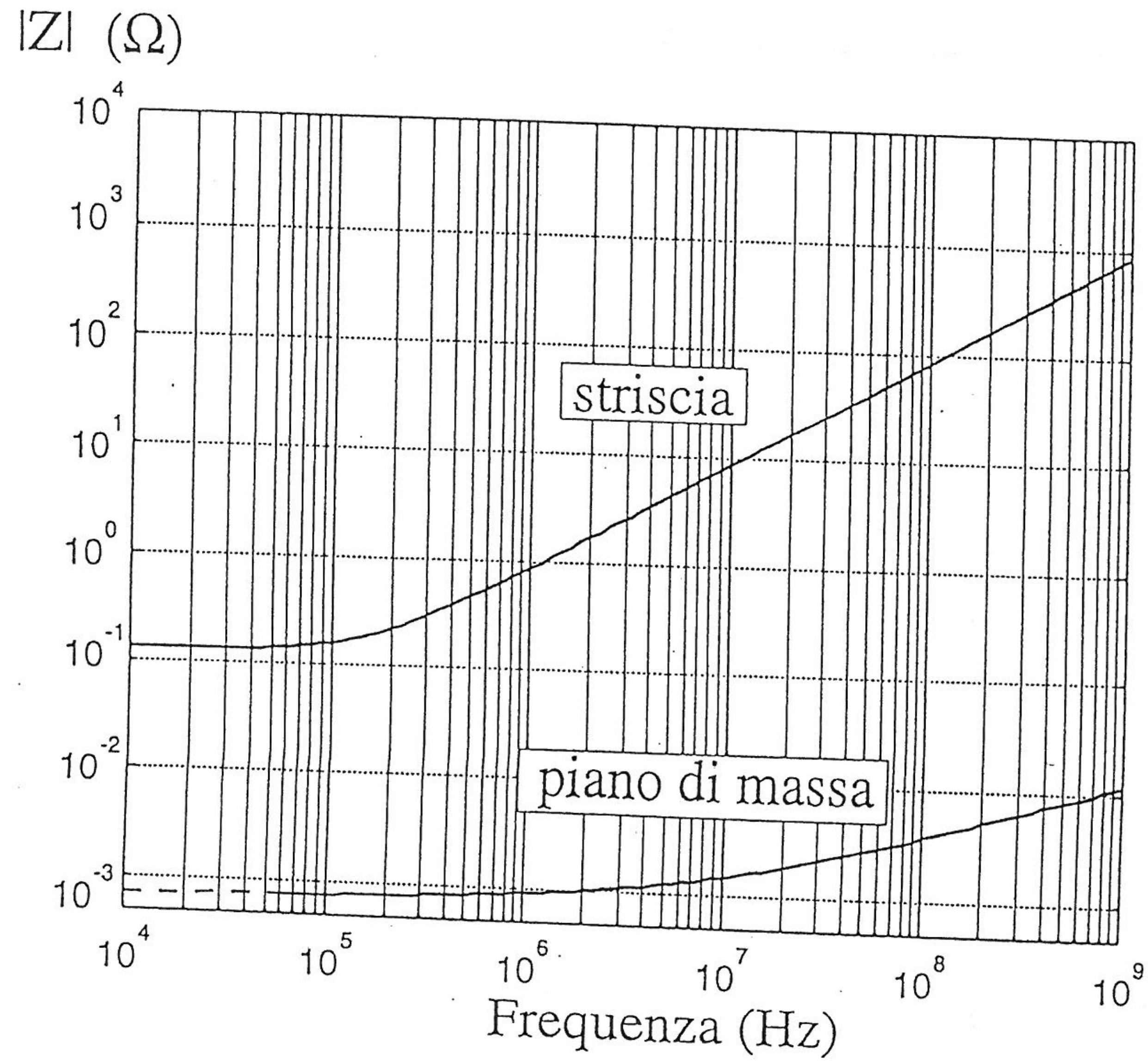
CASO PARTICOLARE PER:

$R = 10\ k\Omega$ $\quad$ $C_P = 0.5\ pF$

$C = 100\ pF$ $\quad$ $L_P = 90\ nH$

$R_P = 100\ m\Omega$

CONFRONTO IMPEDENZA STRISCIA — PIANO DI MASSA



CONDUTTORE : RAME

STRISCIA :

$W = 15\ mil$

$\ell = 10\ cm$

$t = 35\ \mu m$

PIANO DI MASSA :

$t = 35\ \mu m$

CORREZIONE PER

t NON $\gg \delta$

Figure estratte dalla seguente nota applicativa:

ABCs of Probes

Primer

Tektronix

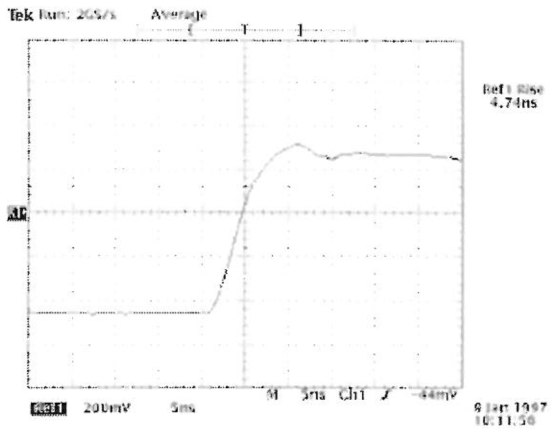


Figure 1.12a. Direct probe tip contact.

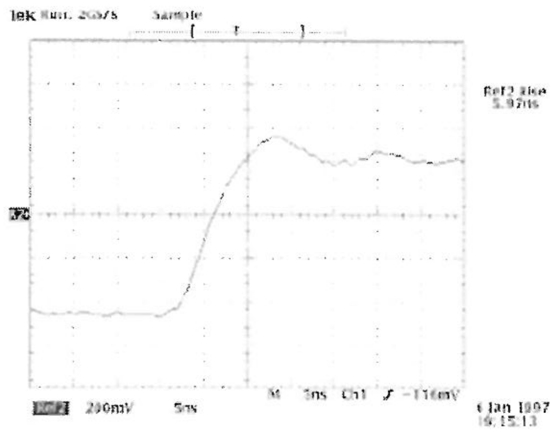


Figure 1.12b. Two-inch wire at probe tip.

Figure 1.12a and Figure 1.12b. Even a short piece of wire soldered to a test point can cause signal fidelity problems. In this case, rise time has been changed from 4.74 ns (Figure 1.12a) to 5.67 ns (Figure 1.12b)

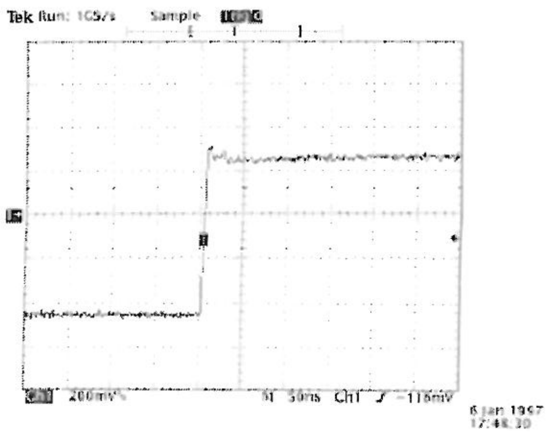


Figure 1.13a. 6.5 inch probe ground lead.

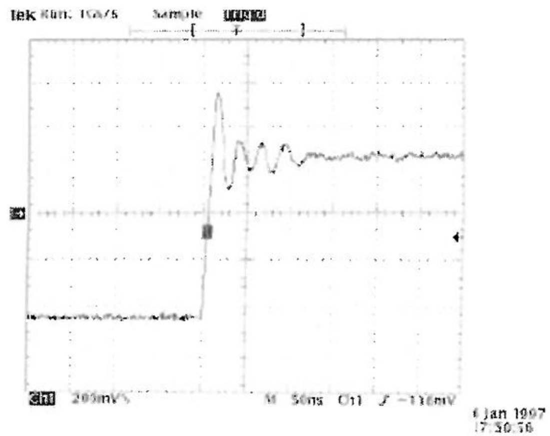
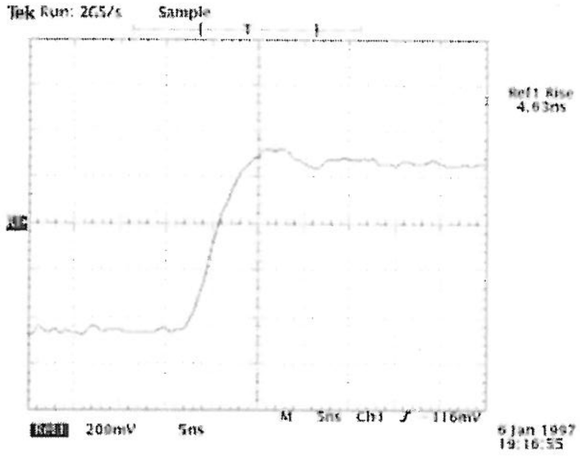
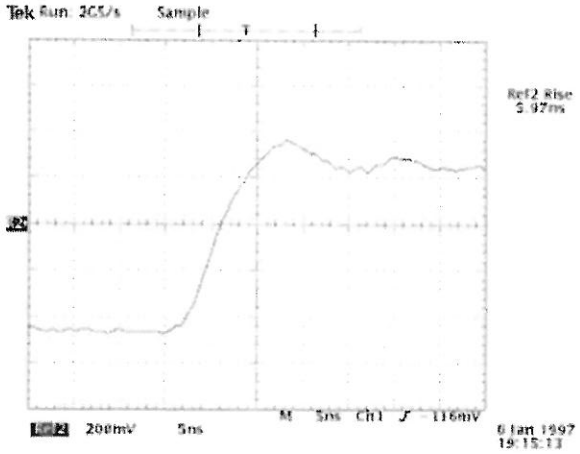


Figure 1.13b. 28 inch lead attached to probe lead.

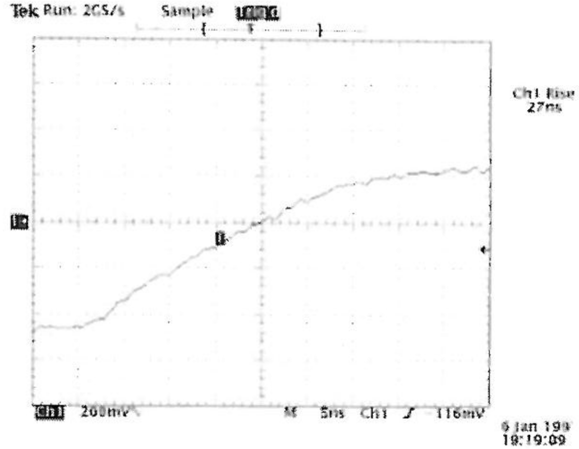
Figure 1.13a and Figure 1.13b. Extending the length of the probe ground lead can cause ringing to appear on pulses.



a.

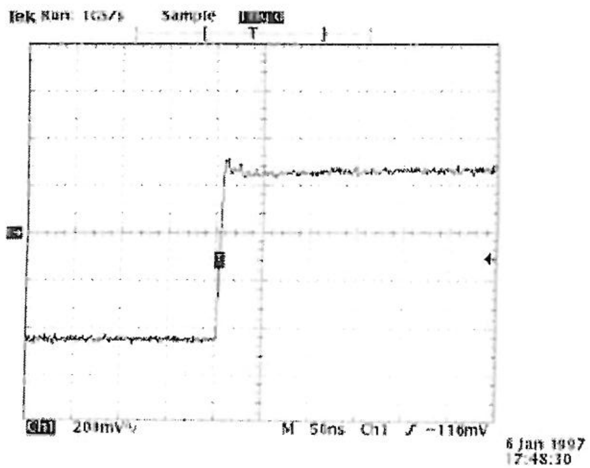


b.

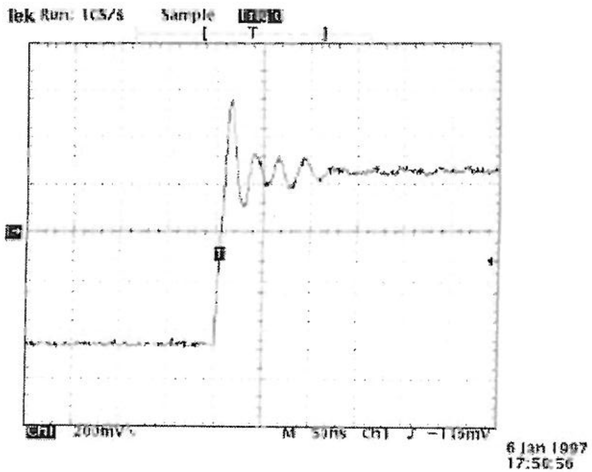


c.

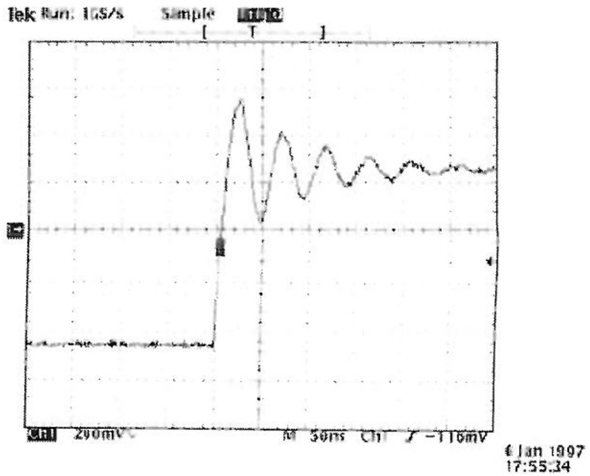
Figure 4.9a, Figure 4.9b, and Figure 4.9c. Effects on rise time of three different probes: (a) 400 MHz, 10X probe, (b) 100 MHz, 10X probe, and (c) 10 MHz, 1X probe. All measurements were made with the same 400 MHz oscilloscope.



a.



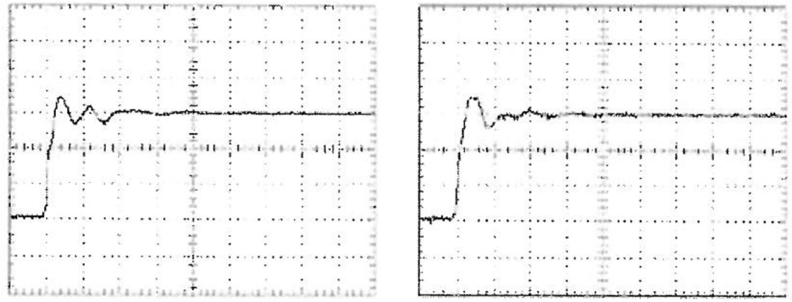
b.



c.

In queste figure i casi a. e b. sono identici a 1.13a e 1.13b. Nel caso c. il conduttore di massa è staccato dalla sonda ed un collegamento di 28 pollici va dalla massa del circuito allo chassis dell'oscilloscopio (oscilloscopio di banda 400 MHz).

Banda oscilloscopio: 350 MHz



a. b.

Figure 6.1a and Figure 6.1b. A fast step (1 ns Tr) has aberrations impressed on it due to use of a six-inch probe ground lead (a). These aberrations can be changed by moving the probe cable or placing a hand over the cable (b).

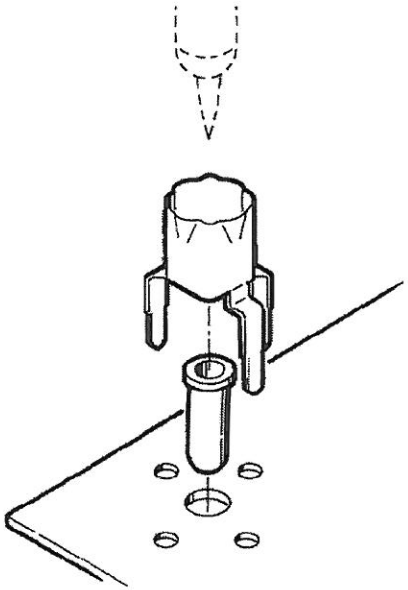


Figure 6.2. Typical ECB-to-Probe Tip Adapter installation

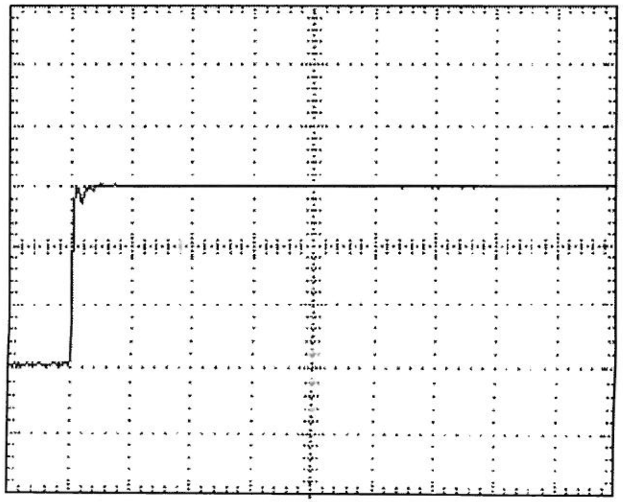


Figure 6.3. The 1 ns rise time step waveform as acquired through an ECB-to-Probe Tip Adapter.

SECTION III

OPERATING INFORMATION

3-1. INTRODUCTION.

3-2. The Model 432A Power Meter operates with HP temperature-compensated thermistor mounts such as the 8478B and 478A Coaxial, and 486A Waveguide series. The frequency range of the 432A with these mounts in 50-ohm coaxial systems is 10 MHz to 18 GHz; in waveguide systems it is 2.6 GHz to 40 GHz. Full-scale power ranges are 10 microwatts to 10 milliwatts (-20 dBm to +10 dBm). Extended measurements may be made to 1 microwatt (-30 dBm). The total measurement capacity of the instrument is divided into seven ranges, selected by a front-panel RANGE switch.

3-3. This section describes general operating procedures and error analysis in microwave power measurement. Application Note 64, available on request from Hewlett-Packard, is a detailed analysis of microwave power measurement problems and techniques.

3-4. CONTROLS, CONNECTORS, AND INDICATORS.

3-4. The front and rear panel controls, connectors, and indicators are explained in Figure 3-2. The descriptions are keyed to the corresponding items which are indicated on the figure.

3-6. The COARSE ZERO and FINE ZERO controls zero the meter. Zero carry-over from the most sensitive range to the other six ranges is within ±0.5%. When the RANGE switch is set to COARSE ZERO, the meter indicates thermistor bridge unbalance, and the front panel COARSE ZERO adjust is for initial bridge balance. For best results, FINE ZERO the 432A on the particular meter range in use.

3-7. The CALIBRATION FACTOR switch provides discrete amounts of compensation for measurement uncertainties related to SWR and thermistor mount efficiency. The Calibration Factor value permits direct meter reading of the RF Power delivered to an impedance equal to the characteristic impedance (Z_O) of the transmission line between the thermistor mount and the RF source. Calibration Factor values are marked on the label of each 8478B, 478A or 486A Thermistor Mount. For further details, see Paragraph 3-23.

3-8. The MOUNT RESISTANCE switch on the front panel compensates for three types of thermistor mounts. Model 486A waveguide mounts can be used by setting the MOUNT RESISTANCE switch to 100Ω or 200Ω, depending on the thermistor mount used (refer to Table 1-2). The 200Ω position is used with Models 478A and 8478B Thermistor Mounts.

3-9. The rear-panel BNC connected labeled RECORDER provides an output voltage linearly proportional to the meter current; 1 volt into an open circuit equals full-scale meter deflection. This voltage is developed across a 1K resistor; therefore, when a recorder with a 1K input impedance is connected to the RECORDER output, approximately .5 volt will equal full scale deflection. This loading of the RECORDER output has no effect on the accuracy of the 432A panel meter.

3-10. A digital voltmeter can be connected to the rear panel RECORDER output for more resolution of power meter readings. When a voltmeter with input impedance greater than 1 megohm is connected to the RECORDER output, 1 volt equals full scale deflection.

3-11. The 432A has two calibration jacks (V_{RF} and V_{COMP}) on the rear panel that can be used for precision power measurements. Instrument error can be reduced from ±1% to ±(0.2% of reading +5μW) of reading, depending on the care taken in measurement and on the accuracy of auxiliary equipment. For further information, see Paragraph 3-27.

3-12. BATTERY OPERATION.

3-13. The Model 432A Option 001 instruments contain tery and conventional 115- or 230-volt line power. A rechargeable Nickel-Cadmium battery is factory-installed in Option 01 instruments. The same battery can be ordered and later installed on the basic instrument, thereby modifying the power meter to the Option 01 configuration. The battery installation kit, HP part number 00432-6016 (including battery charging circuit) may be ordered from the nearest HP Sales Office.

3-14. It is recommended that the Model 432A be battery-operated for up to eight hours, and then allowed to recharge eight hours, or overnight. Continuous battery operation is possible for up to about 20 hours, but then the battery must be recharged for about 20 hours.

3-15. The 432A automatically operates on its internal battery whenever the ac line power is disconnected and the POWER switch is ON. When the battery terminal voltage decreases far enough to force the power supply voltage regulator out of regulation, then the meter stops working and the meter indicator points to the red RECHG BAT. To recharge the battery, simply connect the 432A to ac line power, and turn it ON.

3-16. Battery Storage. Storage of the battery at or below room temperature is best. Extended storage at temperatures above room temperature will reduce cell charge, but will not damage the battery; however, the battery should not be stored where the temperature exceeds 60°C (+140°F).

3-17. MICROWAVE POWER MEASUREMENT ACCURACY.

3-18. A number of factors affect the overall accuracy of power measurement. The major sources of error are mismatch error, RF losses, and instrumentation error.

3-19. Mismatch Error. In a practical measurement situation, both the source and thermistor mount have SWR, and the source is seldom matched to the thermistor mount unless a tuner is used. The amount of mismatch loss in any measurement depends on the total SWR present. The impedance that the source sees is determined by the acutal thermistor mount impedance, the electrical length of the line, and the characteristic impedance of the line, Z_o.

3-20. In general, neither the source nor the thermistor mount has Z_o impedance, and the actual impedances are known only as reflection coefficients, mismatch losses, or SWR. The power delivered to the thermistor mount - and hence the mismatch loss - can only be described as being somewhere between two limits. The uncertainty of power measurement due to mismatch loss increases with SWR. Limits of mismatch loss are generally determined by means of a chart such as the Mismatch Loss Limits charts in Application Note 64. The total mismatch loss uncertainty in power measurement is determined by algebraically adding the thermistor mount losses to the uncertainty caused by source and thermistor mount Z_o match.

3-21. RF Losses. RF losses account for the power entering the thermistor mount but not dissipated in the detection thermistor element. Such losses may be in the walls of a waveguide mount, the center conductor of a coaxial mount, capacitor dielectric, poor connections within the mount, or due to radiation.

3-22. Instrumentation Error. The degree of inability of the instrument to measure the substitution power supplied to the thermistor mount is called power meter accuracy or instrumentation error. Instrumentation error of the Model 432A is ±1% of full scale, 0°C to +55°C.

3-23. CALIBRATION FACTOR AND EFFECTIVE EFFICIENCY.

3-24. Calibration factor and effective efficiency are correction factors for improving power measurement accuracy. Both factors are marked on every HP thermistor mount. Calibration factor compensates for thermistor mount VSWR and RF losses whenever the thermistor mount is connected to an RF source without a tuner. Effective efficiency compensates for thermistor mount RF losses when a tuner is used in the measurement system.

3-25. When the 432A CALIBRATION FACTOR selector is set to the appropriate factor indicated on the thermistor mount, the power indicated by the meter is the power that would be delivered by the source to a load impedance equal to Z_o. More accurately, the relationship between indicated power and the power available to a Z_o load is given by the following equation:

$$P_o = \frac{P\ \text{indicated}\ (1 \pm \rho_s \rho_m)^2}{\text{Calibration Factor}}$$

where

P_o = power available to a Z_o load

ρ_s = source reflection coefficient

ρ_m = thermistor mount reflection coefficient

$$\rho = \frac{SWR - 1}{SWR + 1}$$

Calibration factor does not compensate for source VSWR, or for multiple reflections between the source and the thermistor mount.

3-26. To minimize mismatch between the source and the thermistor mount without a tuner, insert a low SWR precision attenuator in the transmission line between the thermistor mount and the source. Since the mount impedance (and corresponding SWR) deviates significnatly only at the high and low ends of a microwave band, it is generally unnecessary to use a tuner. A tuner or other effective means of reducing mismatch error is recommended when the source SWR is high or when more accuracy is required. For further details, there is a complete discussion of microwave power measurement with emphasis on modern techniques, accuracy considerations and sources of error available in Application Note 64.

3-27. PRECISION POWER MEASUREMENT.

3-28. GENERAL.

3-29. Using precision instruments and careful procedures, measurement error can be reduced to ±0.2% of reading +0.5 μW. The technique involves: 1) zeroing the bridge circuits and measuring the bridge amplifier output voltage difference with a digital voltmeter, then 2) connecting RF power to the thermistor mount and then measuring the bridge amplifier output voltage difference again, and 3) calculating the power from the two measurements. Figure 3-1 shows the instrument setup for dc substitution measurement. Use an HP Model 3440A DVM, with a 3443A Plug-in Unit or a digital voltmeter with equivalent accuracy.

3-30. MEASUREMENT PROCEDURE.

a. Connect the DVM to the 432A rear panel V_{comp} and V_{RF} outputs. Be sure that the digital voltmeter input is isolated from chasses ground.

b. Turn off, or disconnect the RF power from the thermistor mount.

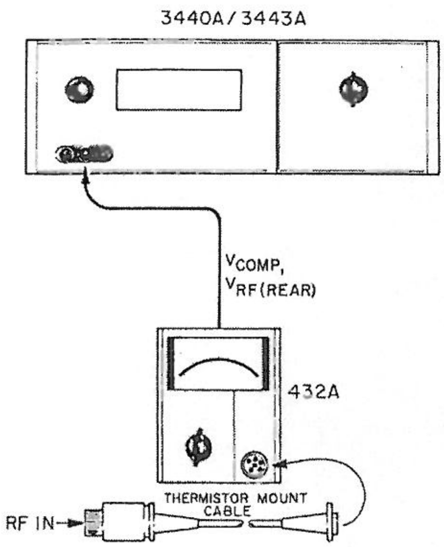


Figure 3-1. Precision Power Measurements

c. Zero the 432A with the COARSE ZERO controls.

d. Depress the FINE ZERO toggle, and measure the differential voltage (V_0) between V_{comp} and V_{RF}.

$$V_0 = V_{COMP} - V_{RF}$$

e. Release the FINE ZERO toggle, and turn on, or reconnect the RF power to the thermistor mount.

f. Measure again the differential voltage (V_1) between V_{RF} and V_{COMP}.

$$V_1 = V_{COMP} - V_{RF}$$

g. Measure V_{COMP} to ground.

h. Calculate incident RF power from the equation

$$P_{RF} = \frac{\frac{1}{4R}\left[2\,V_{COMP}(V_1 - V_0) + V_0^2 - V_1^2\right]}{\text{EFFECTIVE EFFICIENCY}}$$

where
R is the thermistor mount resistance.

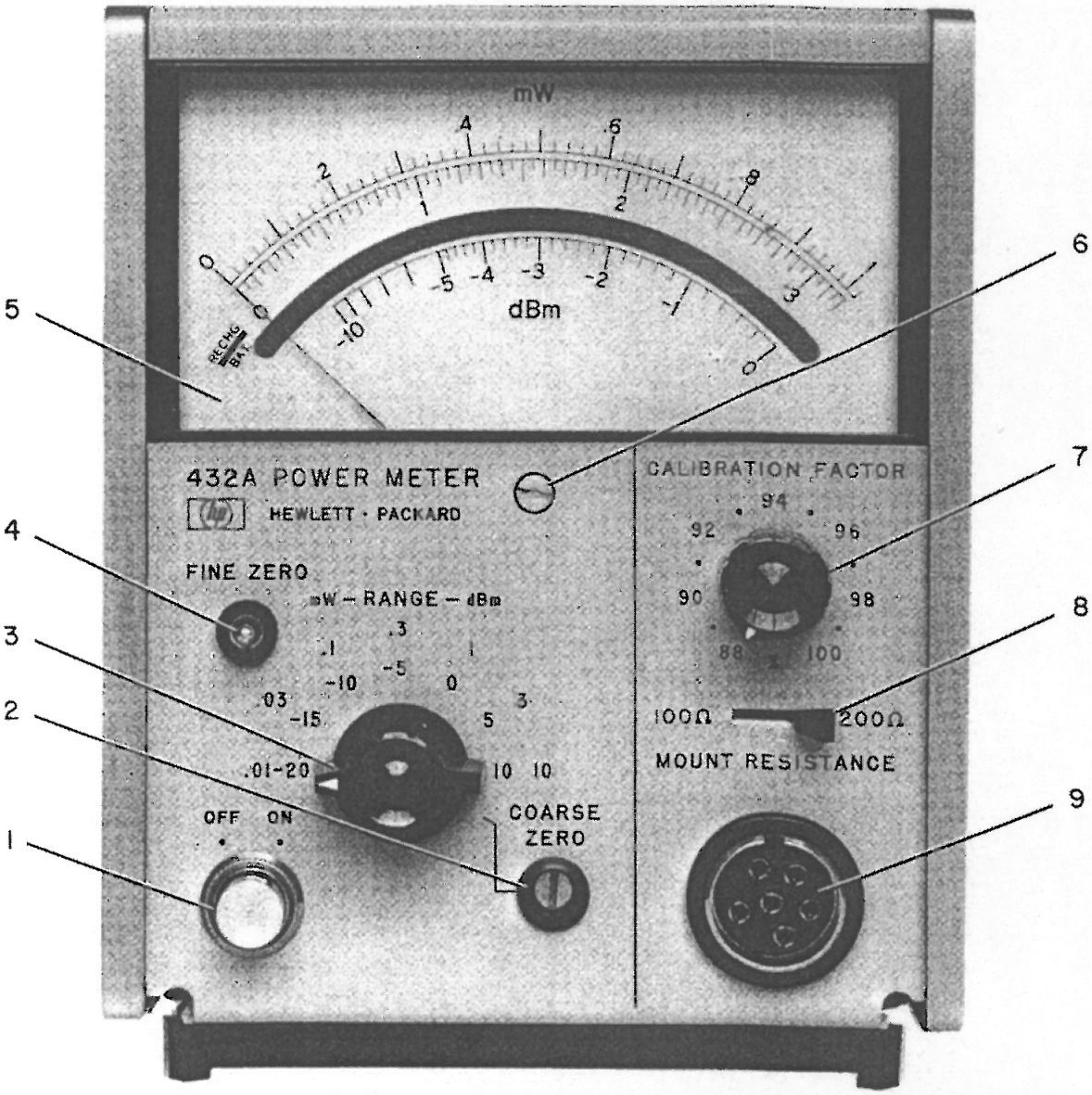


Figure 3-2. Front Panel Controls, Connectors and Indicators (Sheet 1 of 2)

1. POWER. Instrument power ON/OFF switch; connects either ac line voltage or internal battery (Option 01 only) to internal voltage regulator circuits. When ac power is on, optional battery charging circuit operates.

2. COARSE ZERO. Meter zero adjustment; set the RANGE selector to COARSE ZERO, turn OFF the RF power, and adjust to zero the meter.

3. RANGE. Power measurement range selector; selects ranges from 0.01 to 10 milliwatts (-20 to +10 dBm). COARSE ZERO setting is used to zero meter with no power applied to thermistor mount.

4. FINE ZERO. Electronic zero that balances the compensation bridge with zero RF input. To zero meter during operation, close the switch momentarily. Be sure that RF power is not applied to the thermistor mount when the FINE ZERO switch is depressed.

5. Meter. Indicates power input to thermistor mount in milliwatts and dBm. To use the dBm scale, note the value in dBm of the range in use, and subtract from it the reading on the meter dBm scale.

6. Mechanical Meter Zero. Sets meter suspension so that meter indicates zero. To adjust the zero:

   a. Turn POWER switch off.

   b. Turn the adjustment screw clockwise until the indicator falls below zero and comes back up to zero again.

   c. Turn the adjustment very slightly counterclockwise to free up the mechanism from the adjusting peg.

7. CALIBRATION FACTOR. Amplifier gain compensation selector. Set to correspond to the calibration factor printed on the thermistor mount body. See Paragraph 3-23 for more information.

8. MOUNT RESISTANCE. Selects resistance equal to that of mount in use to balance bridges. Table 1-2 lists Hewlett-Packard thermistor mounts and resistances. Set with meter power OFF, when mount is initially connected to the meter.

9. Thermistor Mount Cable Connector. Input connector for 5-1/2 foot cable that connects to the 478A, 8478B, or 486A Thermistor Mounts.

Figure 3-2. Front Panel Controls, Connectors and Indicators (Sheet 2 of 2)

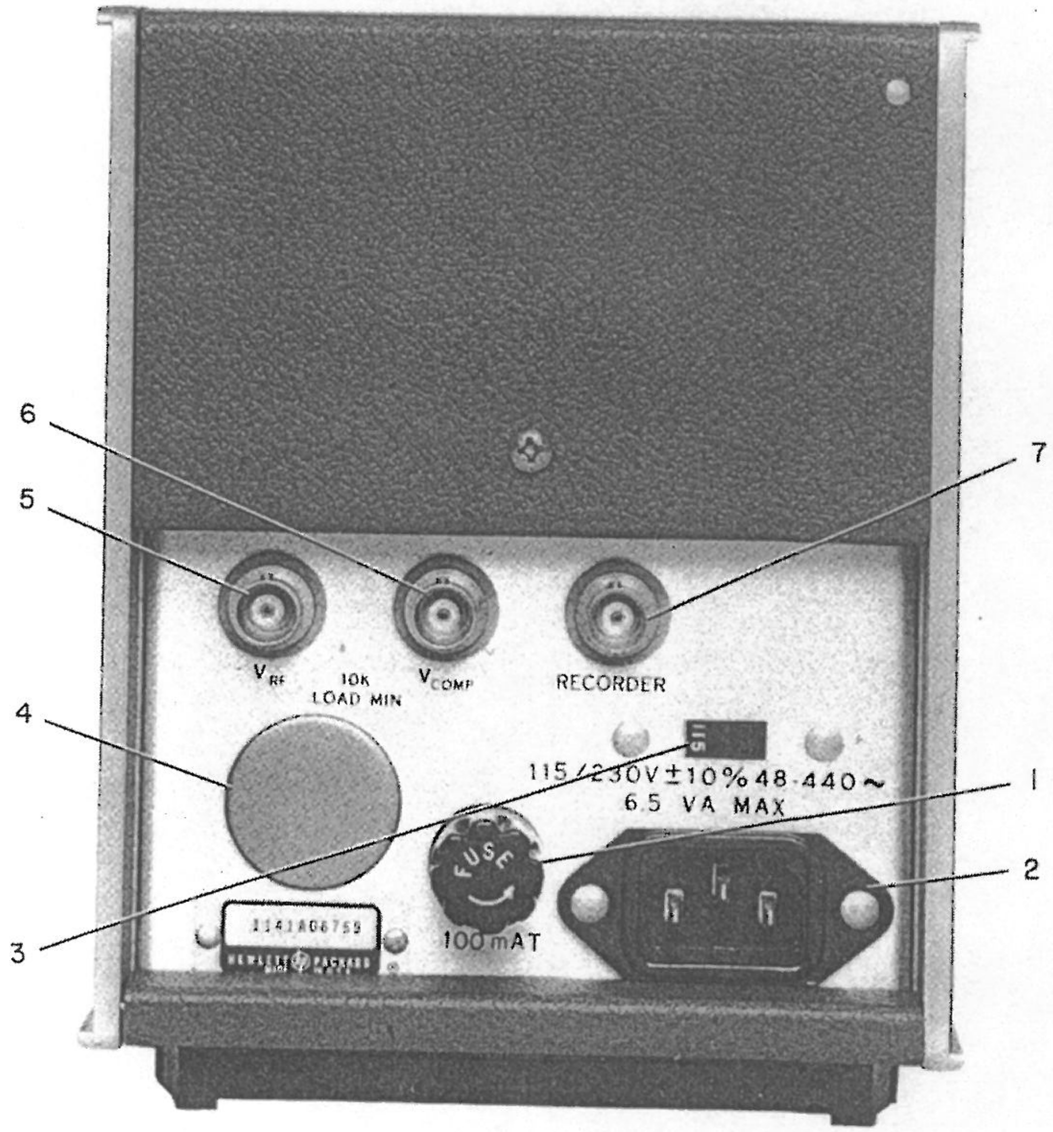


Figure 3-3. Rear Panel Controls and Connectors (Sheet 1 of 2)

1. Line Fuse. For 115 Vac or for 230 Vac use 1/8 amp fuse.

2. Power Cord Input. Use power cord provided, HP 8120-0078. Line power limits are 115/230 Vac, 48-440 Hz. Check FUSE rating and position of line voltage slide switch before connecting power.

3. Line Voltage Slide Switch: Set to line voltage available (115 or 230 Vac, 48-440 Hz).

4. Mounting Hole for Option 002 Model Power Meters. Thermistor mount cable connector installed and wired in parallel with front-panel connector. Only one mount at a time may be used with the power meter.

5. V_{RF} Input. Connected directly to RF bridge. Used for calibrating power meter with HP 8477A Power Meter Calibrator. Also used for precision power measurements.

6. V_{COMP} Input. Connected directly to compensation bridge. Used for calibrating power meter with HP 8477A Power Meter Calibrator. Also used for precision power measurements.

7. RECORDER OUTPUT. Voltage from meter circuit to be used for recorder or digital voltmeter. Output impedance is approx. 1000Ω.

Figure 3-3. Rear Panel Controls and Connectors (Sheet 2 of 2)

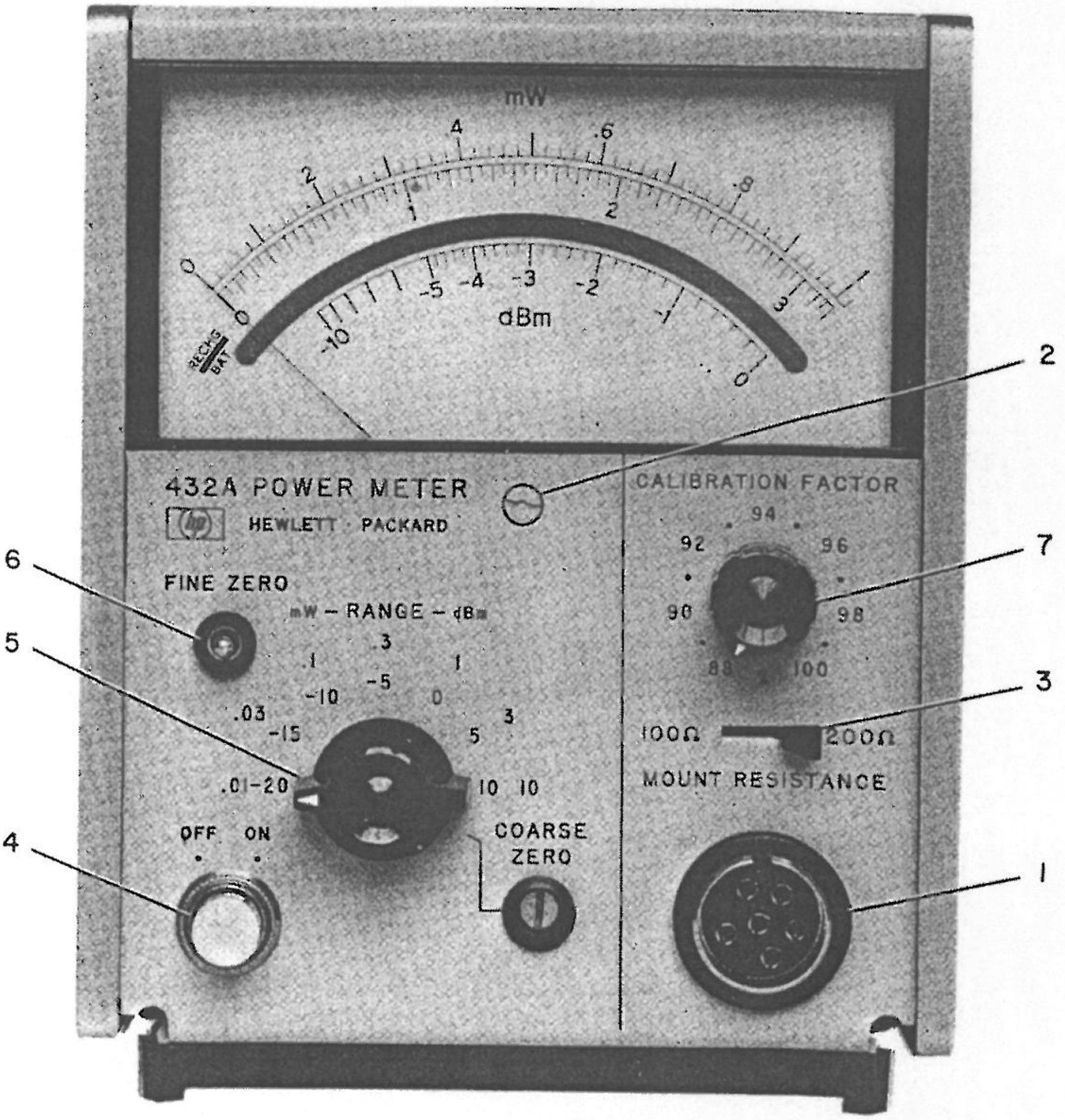


Figure 3-4. Turn On and Zeroing Procedure (Sheet 1 of 2)

1. Connect the thermistor mount and cable to THERMISTOR MOUNT connector. Refer to Table 1-2 for recommended thermistor mounts and their frequency ranges.

2. Meter Mechanical Zero:

   a. With the instrument turned off, rotate the meter adjustment screw clockwise until the pointer approaches the zero mark from the left.

   b. Continue the clockwise rotation until the pointer coincides with the zero mark. If the pointer overshoots, continue rotating the adjustment screw clockwise until the pointer once again approaches the zero mark from the left.

   c. Rotate the adjustment screw about three degrees counterclockwise to disengage screw adjustment from the meter suspension.

3. Set the MOUNT RES switch to correspond to the operating resistance of thermistor mount used.

4. Turn the 432A POWER switch ON. For battery operation, the AC LINE indicator does not turn on.

5. Set the RANGE selector to COARSE ZERO and then zero the meter with the COARSE ZERO screwdriver adjustment.

   Note

   The power meter should be zeroed with the RF power source turned off, or the mount disconnected from the source.

6. Set the range selector to the 0.01 mW range; then depress the FINE ZERO switch until the meter indicates zero.

   Note

   Range-to-range zero carryover is less than ±0.5% if the meter zero has been adjusted (step 2 above), and the instrument has been properly zero-set on the sensitive range. For maximum accuracy, zero-set the power meter on the range to be used.

7. Set CALIB FACTOR switch to correspond to Calibration Factor imprinted on HP thermistor mount label.

8. Apply RF power to the thermistor mount. Power is indicated on the meter directly in mW or dBm.

Figure 3-4. Turn On and Zeroing Procedure

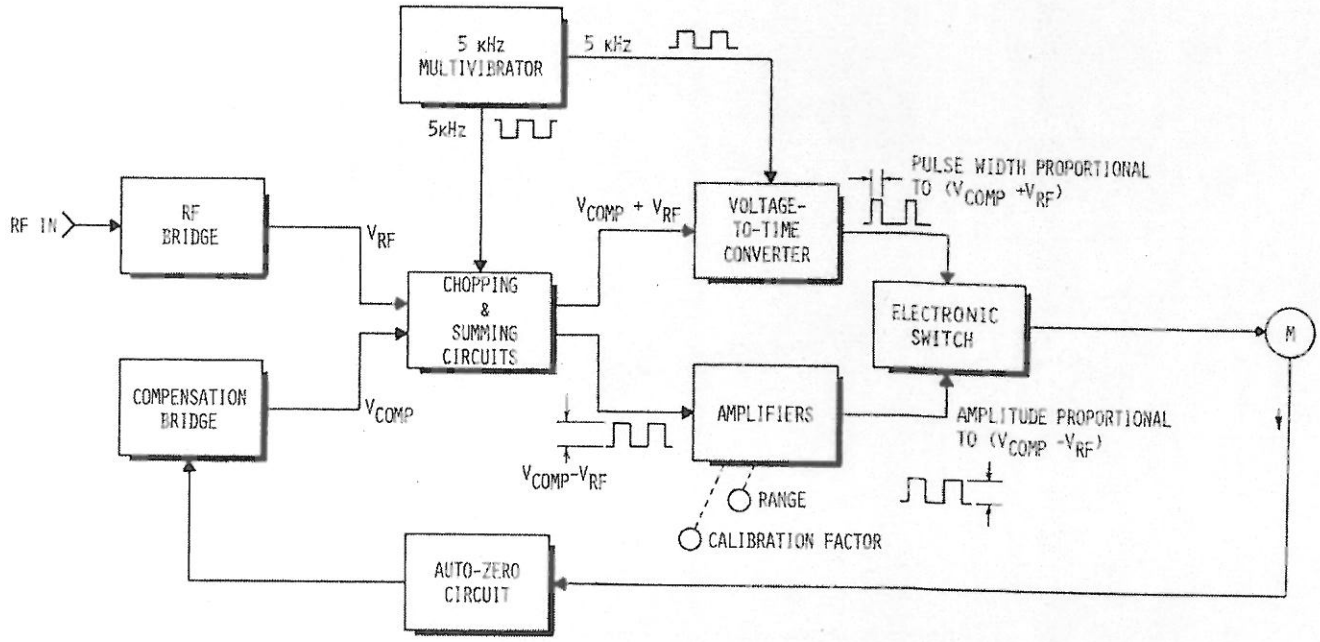


Figure 4-1. Simplified 432A Block Diagram

SECTION IV
PRINCIPLES OF OPERATION

4-1. SIMPLIFIED DESCRIPTION

4-2. The HP 432A Power Meter consists of two major sections: the bridge and meter logic assemblies. The instrument also contains an auto zero circuit which provides for automatic zeroing on any range. A simplified block diagram of the HP 432A is shown in Figure 4-1.

4-3. The bridge section contains circuits which form two self-balancing bridge circuits when a suitable thermistor mount is connected to the 432A. Each bridge is automatically brought to balance by the action of a high gain dc amplifier feeding power to the top of the bridge. The voltage at the top of the RF bridge, V_{RF} is responsive to both input RF power and ambient temperature changes. The voltage at the top of the compensation bridge, V_{COMP} is responsive only to ambient temperature changes. Knowing V_{RF} and V_{COMP}, the RF power can be calculated.

4-4. The meter logic section processes V_{RF} and V_{COMP} to produce a meter current proportional to RF power. The sum (V_{RF} + V_{COMP}) controls the width of 5 kHz pulses. The difference (V_{COMP} - V_{RF}) is chopped, amplified and fed to an electronic switch actuated by the controlled width pulses. Therefore, the meter current is pulses of variable height and width with the meter indicating the average current. (This process produces a meter current proportional to (V_{RF} + V_{COMP}) (V_{RF} - V_{COMP}). Paragraph 4-10 explains why this is necessary.

4-5. FUNCTIONAL BLOCK DIAGRAM

4-6. A functional block diagram of the 432A power meter is shown in Figure 4-2. The instrument comprises two major assemblies: bridge assembly A1 and meter logic assembly A2. Auto zero circuit A1A1, which provides for automatic zeroing of the instrument, is included as part of logic assembly A1.

4-7. The thermistor bridges are biased with direct current from the bridge amplifiers. Each bridge amplifier supplies enough heating current to bring the thermistor resistance to 100 or 200 ohms, depending upon the setting of the MOUNT RESISTANCE switch on the 432A. If one of the thermistor bridges is unbalanced due to incorrect thermistor resistance, an error voltage occurs and is amplified by the bridge amplifier. The error voltage is applied to the top of the bridge and changes the power dissipation of the negative temperature coefficient thermistor. The change of power dissipation causes the resistance to the thermistor to change in the direction required to balance the bridge. Application of RF power to the RF bridge heats the thermistor and lowers its resistance. The bridge circuit responds by reducing the dc voltage applied to the top of the bridge thus maintaining bridge balance.

4-8. If ambient temperature causes changes in the thermistor resistance, the bridge circuits respond by applying an error voltage to the bridges to maintain bridge balance. The voltage at the top of the RF bridge is dependent upon both ambient temperature and the RF input. The voltage at the top of the compensation bridge is dependent upon the ambient temperature only. The power meter reading is brought to zero with no applied RF power by making V_{COMP} equal to V_{RF} so (V_{COMP} - V_{RF}) equals zero. Since ambient temperature causes both thermistors to respond similarly, there will be no net difference between the amplifier output voltages. Therefore, any difference in output volrages from the bridges is now due to RF power absorbed by the thermistor mount.

4-9. The RF bridge voltage, V_{RF}, and the compensation bridge voltage, V_{COMP}, contain the "RF power" information. To provide a meter reading proportional to RF power the dc voltages (V_{RF}, V_{COMP}) must be further processed by the meter logic circuits.

4-10. The required processing is derived as follows: P_o is absorbed power needed by the RF thermistor to bring its resistance to R ohms (100 or 200 ohms). P_o consists of two components: RF power and dc power supplied by the 432A. The self-balancing action of the bridge circuit automatically adjusts the dc power so that the total power in the thermistor is P_o. This dc power is related to the voltage V_{RF} at the top of the bridge by $(V_{RF}/2)^2/R$. Thus

$$P_o = \text{RF power} + \text{DC power}$$

$$= \text{RF power} + \frac{V_{RF}^2}{4R}$$

4-11. RF power can be determined by measuring V_{RF} with and without applied RF power and then doing some arithmetic. But this power measuring scheme is neither convenient nor temperature compensated (since P_o changes with temperature). The 432A introduces another thermistor bridge circuit exposed to the same ambient temperature but not RF power. This circuit includes adjustments (COARSE and FINE ZERO) so that the dc voltage V_{COMP} at the top of its bridge can be set equal to V_{RF}. Assuming matched RF and compensation thermistors, V_{RF0} (with no RF power) and V_{COMP} remain equal with ambient temperature fluctuation. They differ only when the RF power to be measured is applied to the RF thermistor. Thus, we have

$$V_{COMP} = V_{RF_0} \quad \text{when RF power} = 0$$

and

$$P_o = 0 + \frac{V_{COMP}^2}{4R}$$

Combining equations, we have:

$$\frac{V_{COMP}^2}{4R} = \text{RF power} + \frac{V_{RF}^2}{4R}$$

or

$$\text{RF power} = \frac{V_{COMP}^2 - V_{RF}^2}{4R}$$

$$= \frac{1}{4R}\left(V_{COMP} + V_{RF}\right)\left(V_{COMP} - V_{RF}\right)$$

4-12. Thus an RF power measurement reduces to setting $V_{COMP} = V_{RF0}$ (with zero RF power) initially, measuring V_{COMP} and V_{RF}, and computing with the above formula. The 432A carries out the computation by forming the indicated sum and difference, performing the multiplication and displaying the result on a meter.

4-13. The meter logic circuits change the two dc voltages to two pulse signals which contain all the RF power information. One of the signals will be a square wave whose amplitude is proportional to V_{COMP} - V_{RF}. The other signal will have a pulse width proportional to $V_{COMP} + V_{RF}$.

4-14. The $V_{COMP} - V_{RF}$ signal is obtained by taking the dc voltage outputs from the A1 assembly and applying them to a chopper circuit. This chopper circuit is driven by a 5-kHz multivibrator. The output of the chopper is a square wave signal whose amplitude is proportional to $V_{COMP} - V_{RF}$. The output of the chopper is coupled to the range amplifier and then to the calibration factor amplifier. The amplification that the signal receives in these two amplifiers depends upon the setting of the RANGE switch and the CALIBRATION FACTOR switch. The output of the calibration factor amplifier is V. This current is fed to the electronic switch. A square wave current with amplitude proportional to $(V_{COMP} - V_{RF})$.

4-15. The $V_{COMP} + V_{RF}$ signal is obtained by taking the two dc voltages from A1 assembly through a summing circuit and feeding this voltage to a voltage-to-time converter. The voltage-to-time converter is driven by a 5-kHz multivibrator. The output of the voltage-to-time converter is a signal whose pulse width is proportional to the sum of $V_{COMP} + V_{RF}$. This signal controls the electronic switch. From the $V_{COMP} - V_{RF}$ and $V_{COMP} + V_{RF}$ inputs, the electronic switch provides a 5-kHz pulse train whose amplitude is proportional to $V_{COMP} - V_{RF}$ and whose pulse width is proportional to $V_{COMP} + V_{RF}$. The pulse width is always 90 msec or less.

4-16. The bias circuit switch and filter provides a zero current reference for the meter circuits. This is accomplished by controlling the dc bias to the first stage of the calibration factor amplifier. This circuit, in effect, restores the dc component to the square wave which has been amplified by ac coupled amplifiers.

4-17. The meter is a 0-1 mA, full-scale meter that has a capacitor across its terminals. The capacitor integrates the output pulses from the current switch so the current into the meter is proportional to the time average of the input pulses. That is, the input current to the meter is proportional to the product of

$$\left(V_{COMP} + V_{RF}\right)\left(V_{COMP} - V_{RF}\right)$$

$$= \left(V_{COMP}\right)^2 - \left(V_{RF}\right)^2$$

4-18. The output from the meter is further filtered so the voltage at the rear panel RECORDER output is suitable for use with either a digital voltmeter or X-Y recorder. The RECORDER output voltage is returned to the compensation bridge through the automatic zero circuit when the FINE ZERO switch is depressed. The automatic zero circuit holds a correction voltage at the input of the compensation bridge amplifier, so when the RF is zero, the meter indication will also be zero.

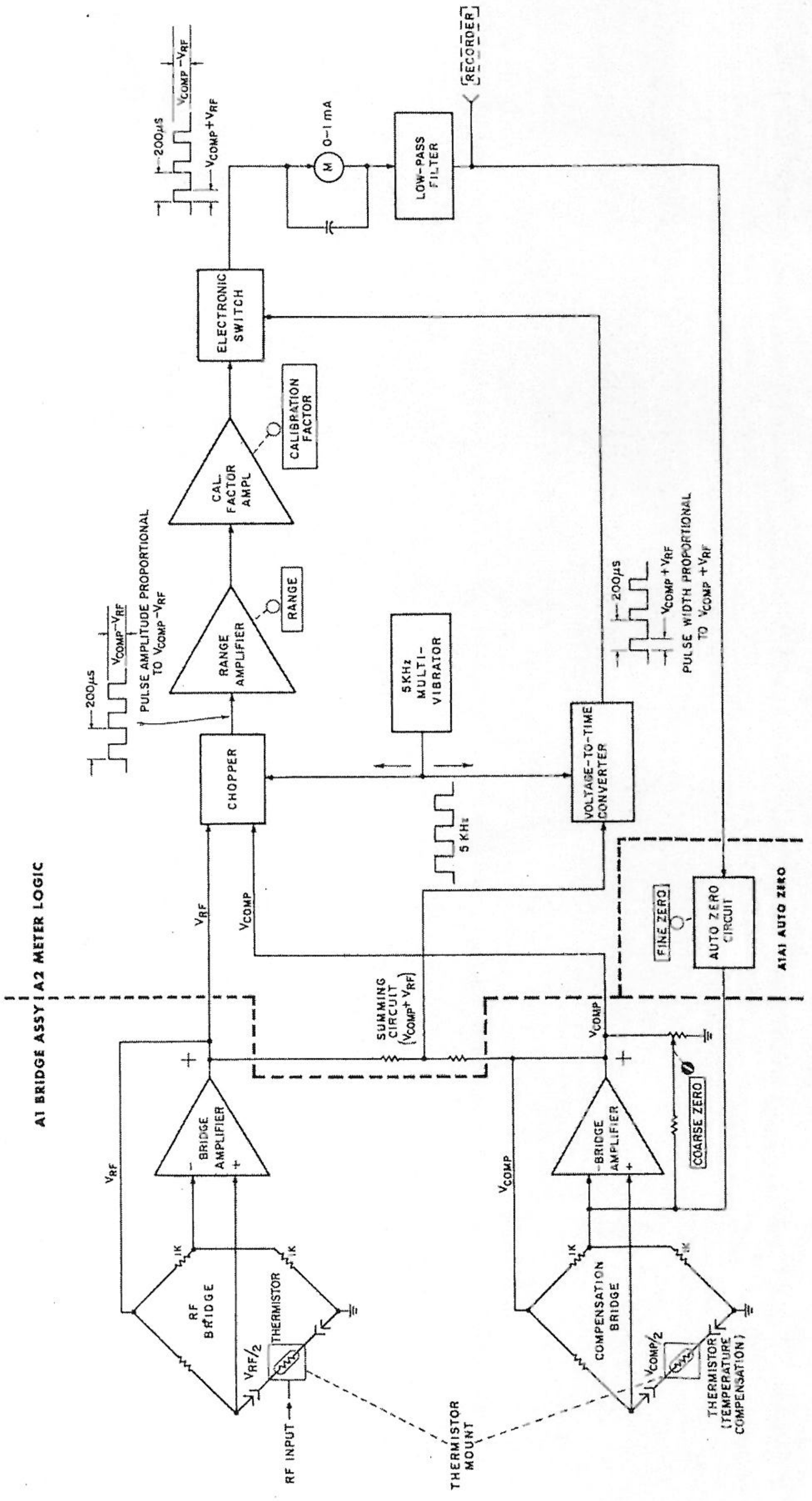


Figure 4-2. Model 432A Block Diagram

Note su

Impedenzimetro Vettoriale

HP 4193A

Carlo Carobbi per l'insegnamento di
Misure Elettroniche
(LM Ingegneria Elettronica-Università degli
Studi di Firenze)

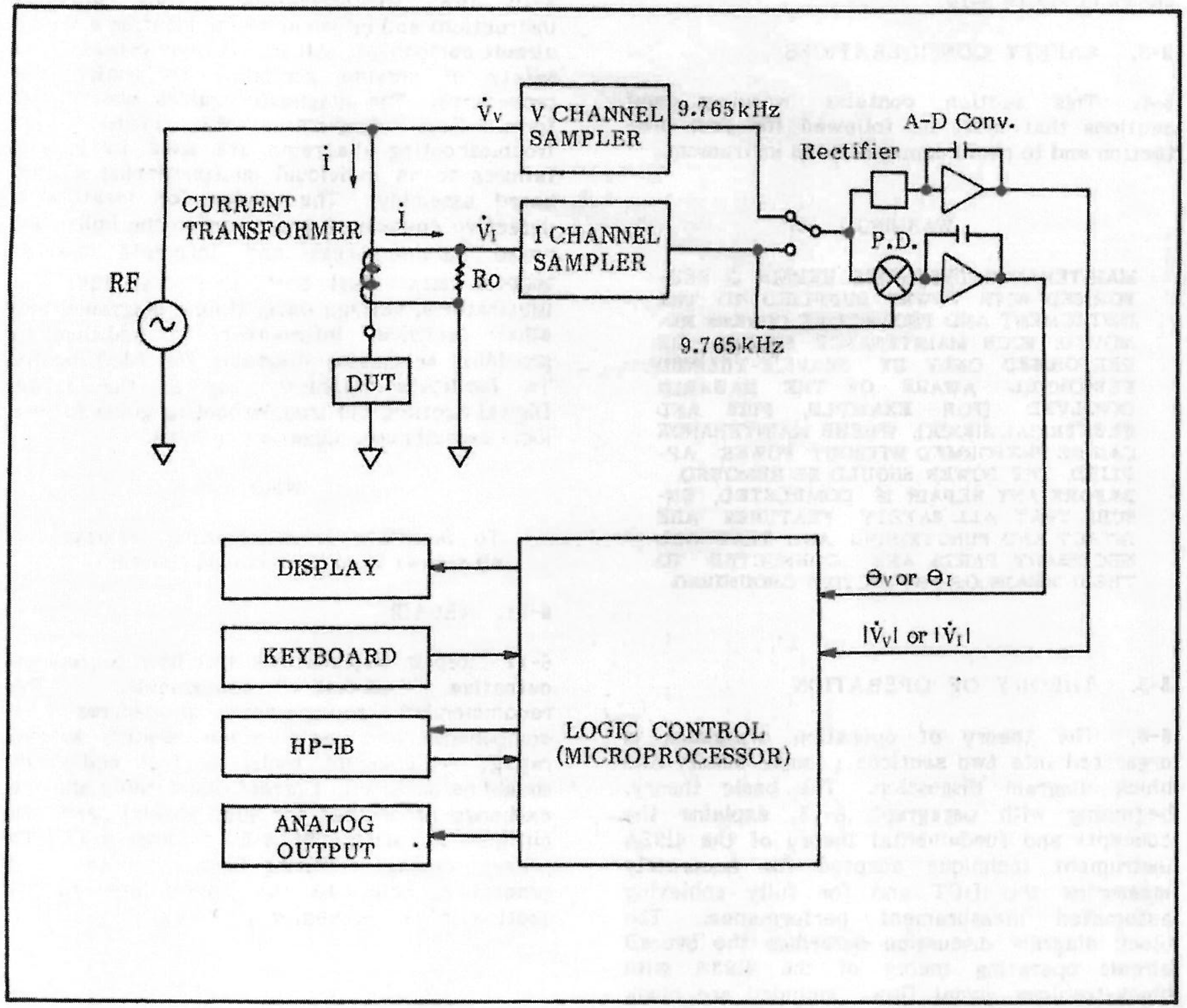


Figure 8-1. Basic Block Diagram.

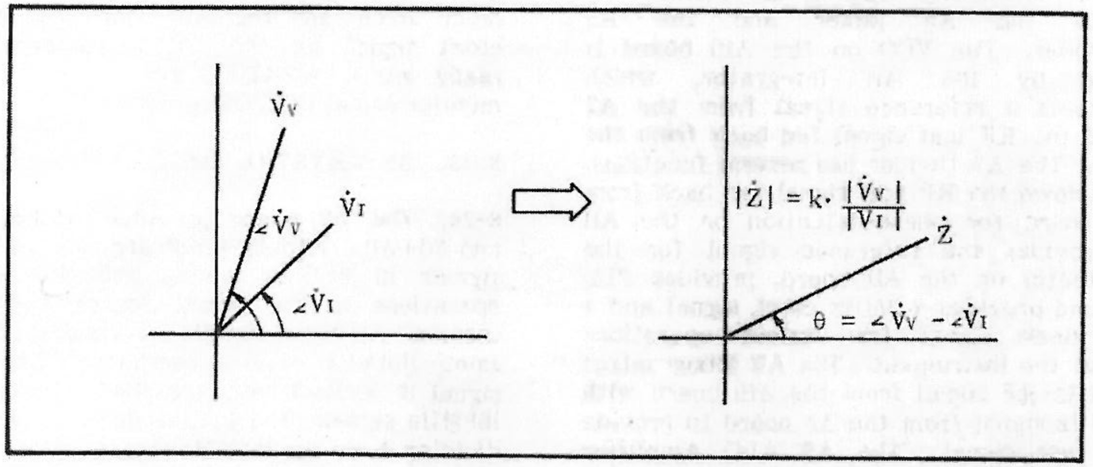


Figure 8-2. Relation between $\dot{Z}$, $\dot{V}_V$, and $\dot{V}_I$.

Conversione di frequenza per tramite di campionamento

Si intende effettuare una conversione di frequenza di un segnale a radiofrequenza (RF). La conversione avviene dalla frequenza del segnale RF, f_{RF}, alla frequenza intermedia (IF), f_{IF}, tramite campionamento a frequenza f_S. Nel caso in esame si ha $f_{IF} \ll f_{RF}$ e f_S comparabile o inferiore a f_{RF}. Il segnale RF è perfettamente ripetitivo.
Si assume che il campionamento consista nell'acquisire una porzione di segnale di durata τ a intervalli di tempo fissi $T_S = 1/f_S$. Poniamo il segnale

$$v(t) = A\cos\left(\omega_{RF}t + \phi\right) \tag{1}$$

ed il segnale campionato

$$v_S(t) = v(t)\cdot s(t), \tag{2}$$

dove $s(t)$ è periodico di periodo T_S e per $0 \le t \le T_S$ è

$$s(t) = \begin{cases} 1 & 0 \le t < \tau \\ 0 & \tau \le t < T_S \end{cases}.$$

Il segnale di campionamento $s(t)$ può essere sviluppato in serie di Fourier, e si trova

$$s(t) = \frac{\tau}{T_S}\sum_{n=-\infty}^{\infty}\frac{\sin\left(n\omega_S\tau/2\right)}{n\omega_S\tau/2}e^{jn\omega_S(t-\tau/2)}\ . \tag{3}$$

La (3) può essere facilmente ricavata tenendo conto che la trasformata di Fourier di un impulso unitario rettangolare singolo di durata τ è

$$S(\omega) = \tau\frac{\sin\left(\omega\tau/2\right)}{\omega\tau/2}e^{-j\omega\tau/2}$$

ed i coefficienti della serie di Fourier dell'impulso ripetuto con cadenza T_S sono dati da $S_n = \dfrac{1}{T_S}S\left(n\omega_S\right)$. Dalla (3) poi si ottiene facilmente

$$s(t) = \frac{\tau}{T_S} + \frac{2\tau}{T_S}\sum_{n=1}^{\infty}\frac{\sin\left(n\omega_S\tau/2\right)}{n\omega_S\tau/2}\cos\left[n\omega_S\left(t-\tau/2\right)\right]. \tag{4}$$

Sostituendo adesso la (1) e la (4) nella (2) si trova, dopo qualche passaggio,

$$
\begin{aligned}
v_S(t) = {} & \frac{A\tau}{T_S}\cos\left(\omega_{RF}t+\phi\right)+ \\
& \frac{A\tau}{T_S}\sum_{n=1}^{\infty}\frac{\sin\left(n\omega_S\tau/2\right)}{n\omega_S\tau/2}\cos\left[\left(n\omega_S+\omega_{RF}\right)t+\phi-n\omega_S\tau/2\right]+ \\
& \frac{A\tau}{T_S}\sum_{n=1}^{\infty}\frac{\sin\left(n\omega_S\tau/2\right)}{n\omega_S\tau/2}\cos\left[\left(n\omega_S-\omega_{RF}\right)t-\phi-n\omega_S\tau/2\right].
\end{aligned}
\tag{5}
$$

Dalla (5) si vede che se si sceglie $f_S = \frac{1}{N}\left(f_{RF}+f_{IF}\right)$ dove N è un intero allora il prodotto di mescolazione corrispondente a $n = N$ nella seconda serie della (5) ha frequenza f_{IF}. L'impedenzimetro vettoriale HP 4193A impiega questa tecnica per convertire alla frequenza intermedia $f_{IF} = 9,765625$ kHz il segnale di tensione e di corrente a frequenza f_{RF}, compresa fra 400 kHz e 110 MHz, proveniente dalla sonda. N è scelto dal microprocessore dell'impedenzimetro a seconda della frequenza f_{RF} impostata dall'utilizzatore ed è compreso fra 1 e 44. La conversione di frequenza non produce alterazioni del rapporto delle ampiezze e della differenza di fase dei segnali di tensione e di corrente. La generazione del segnale di campionamento alla frequenza f_S avviene secondo lo schema in figura

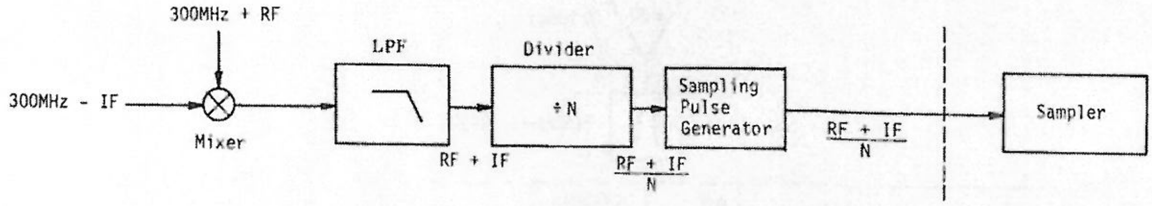


Figure B. Mixing Down Method.

3-41. Probe and Test Fixture Residuals

3-42. The equivalent circuit of the 4193A's measurement port is shown in Figure 3-8. All measured values displayed on the MAGNITUDE and PHASE displays include the residuals of the probe and the test fixture. Typical values of each residual are listed in Table 3-8.

3-43. The conductive component of the open-circuit admittance of the equivalent circuit shown in Figure 3-8 is sufficiently larger than the susceptive component, c, at the frequencies below 110MHz to be negligible.

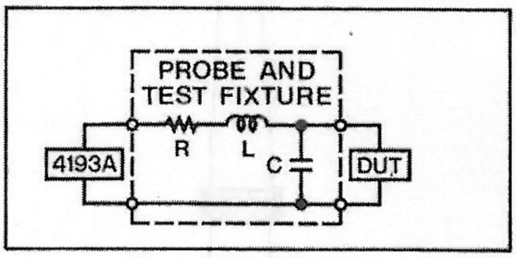


Figure 3-8. Equivalent Circuit.

3-44. Residuals compensation can be made using the following procedure :

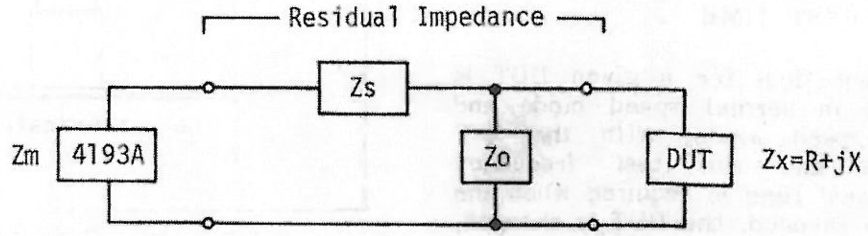
(1). Connect nothing to the test fixture (or probe) and note the value displayed as Zo.

(2) Short the test fixture (or probe) and note the value displayed as Zs.

(3) Calculate the DUT's actual impedance using the equation given in Figure 3-9.

Table 3-8. Typical Residuals at 100MHz

PROBE AND TEST FIXTURE	R (Ω)	L (nH)	C (pF)
PROBE + 16099A	0.5	10	2.4
PROBE + 16099A + 16092A	0.5	11	3.5
PROBE + 16099A + 16093A	0.5	12	4.2
PROBE + 16099A + 16093B	0.5	12	7.9



Zm =measured impedance, Zs =short-circuit impedance,

Zo =open-circuit impedance, Zx =DUT impedance.

$$|Zx| = \sqrt{R^2 + X^2}$$

$$\theta = \tan^{-1} \frac{X}{R}$$

where:

$$R = \frac{(|Z_0| \cos\theta m - |Zm| \cos\theta_0) \cdot |Zm| \cdot |Z_0|}{(|Z_0| \cos\theta m - |Zm| \cos\theta_0)^2 + (|Zm| \sin\theta_0 - |Z_0| \sin\theta m)^2} - |Zs| \cos\theta s$$

$$X = \frac{(|Z_0| \sin\theta m - |Zm| \sin\theta_0) \cdot |Zm| \cdot |Z_0|}{(|Z_0| \cos\theta m - |Zm| \cos\theta_0)^2 + (|Zm| \sin\theta_0 - |Z_0| \sin\theta m)^2} - |Zs| \sin\theta s$$

$|Z_0|$ and θ_0: Open circuit impedance and phase, respectively.

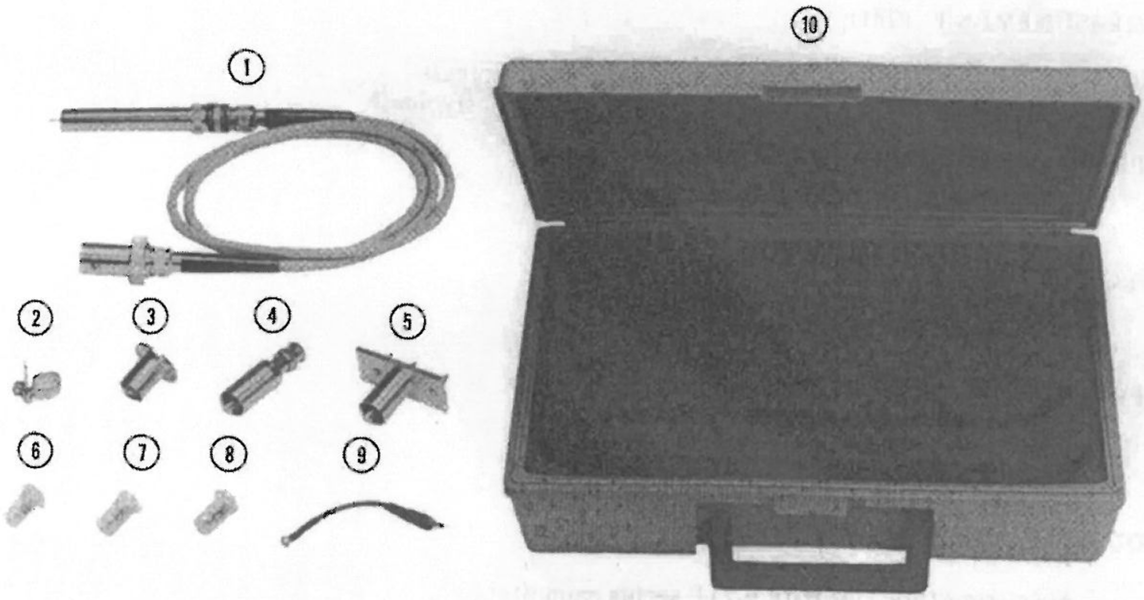
$|Zs|$ and θs: Short circuit impedance and phase, respectively.

Note

These equations assume that $Z_0 >> Zs$.

Figure 3-9. Residuals Compensation.

Table 1-3. Probe Kit for 4193A



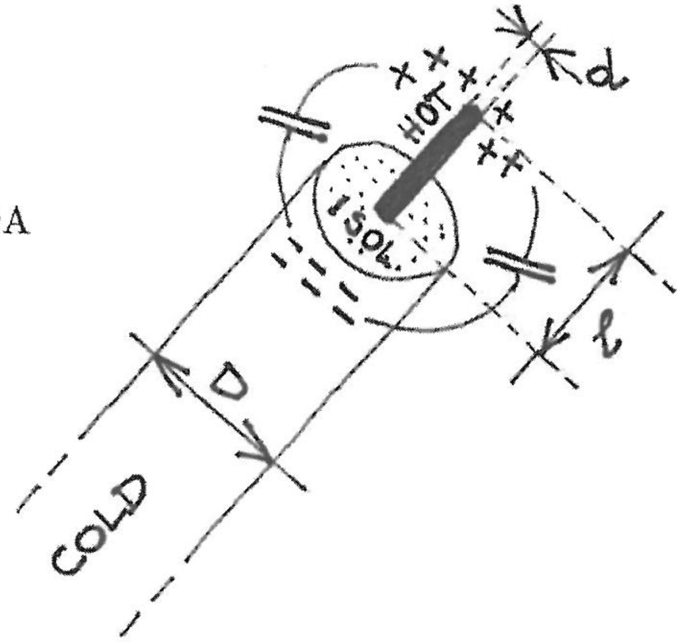
* Kit Part No. : 04193-87001

Reference	HP Part Number	Qty	Description
1	04193-61151	1	PROBE
2	04193-61154	1	GROUND ADAPTER
3	04193-21008	1	PROBE SOCKET
4	04193-61152	1	BNC ADAPTER
5	04193-61153	1	COMPONENT ADAPTER
6	04193-60153	1	SPARE N-TYPE PIN SET Contains five spare N-type pins (HP Part No.: 04193-21023)
7	16095-60012	1	SPARE PIN SET Contains ten spare N-type pins (HP Part No.: 16095-29005)
8	04193-60151	1	SPARE CLIP SET Contains three spare clips (HP Part No.: 0360-2065)
9	04193-61629	1	GROUND LEAD
10	04193-60152	1	PROBE KIT CASE

Parassiti - Capacità di circuito aperto

Residenza fisica, stima, misura (*hp* 4193A)

SONDA



$l = 9.7$ mm

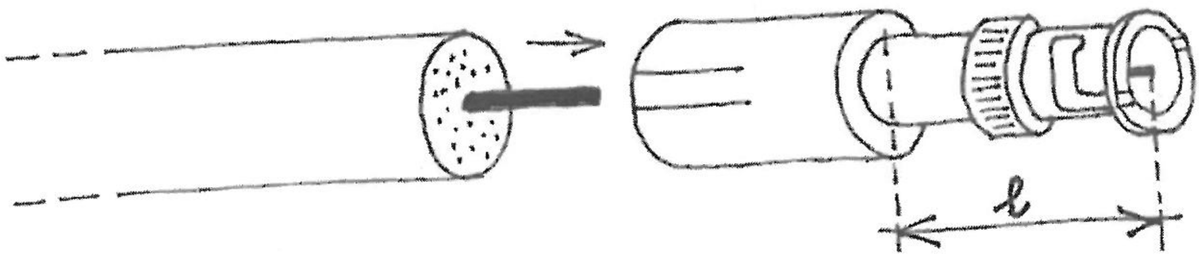
$d = 1.4$ mm

$D = 1.6$ cm

$$C_{os} \approx \frac{2\pi\epsilon_o l}{\ln\left(\frac{2l}{d}\right) - 1} = 0.33 \text{ pF}$$

$C_{om} = 0.22$ pF

SONDA + ADATTATORE BNC



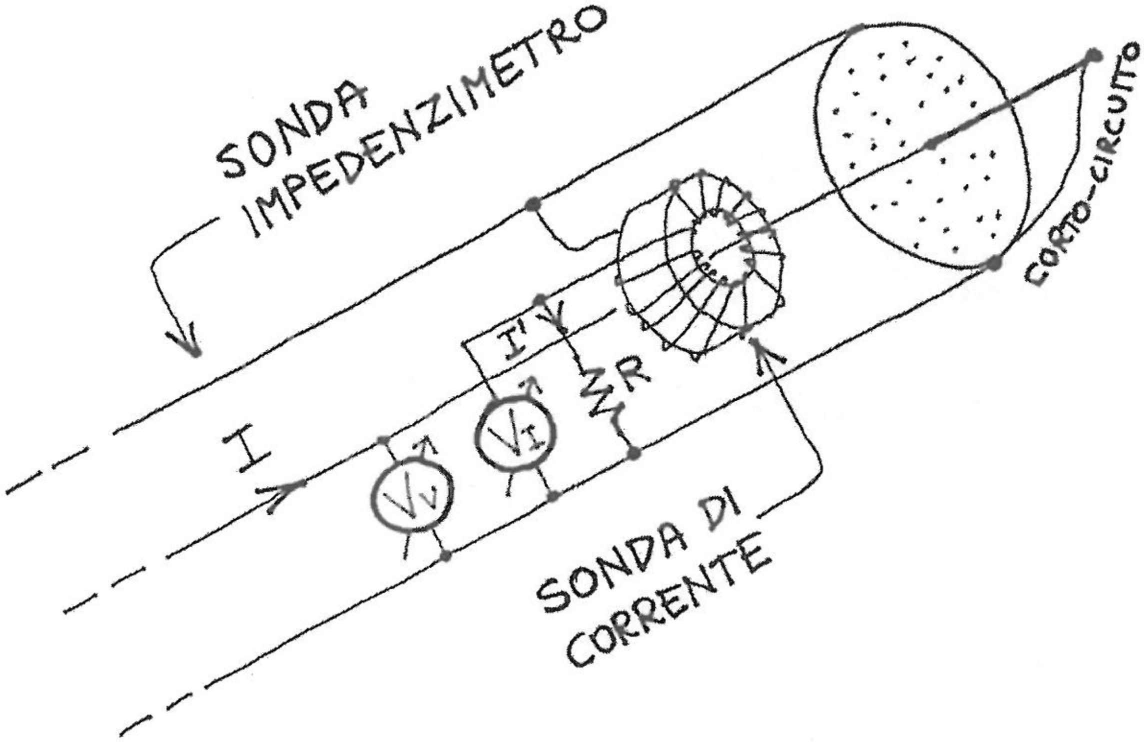
$l = 2.2$ cm

$C_{os} \simeq C_{coax} \cdot l = 2.20$ pF

$C_{om} = 2.47$ pF

Parassiti - Impedenza di corto-circuito (*hp* 4193A)

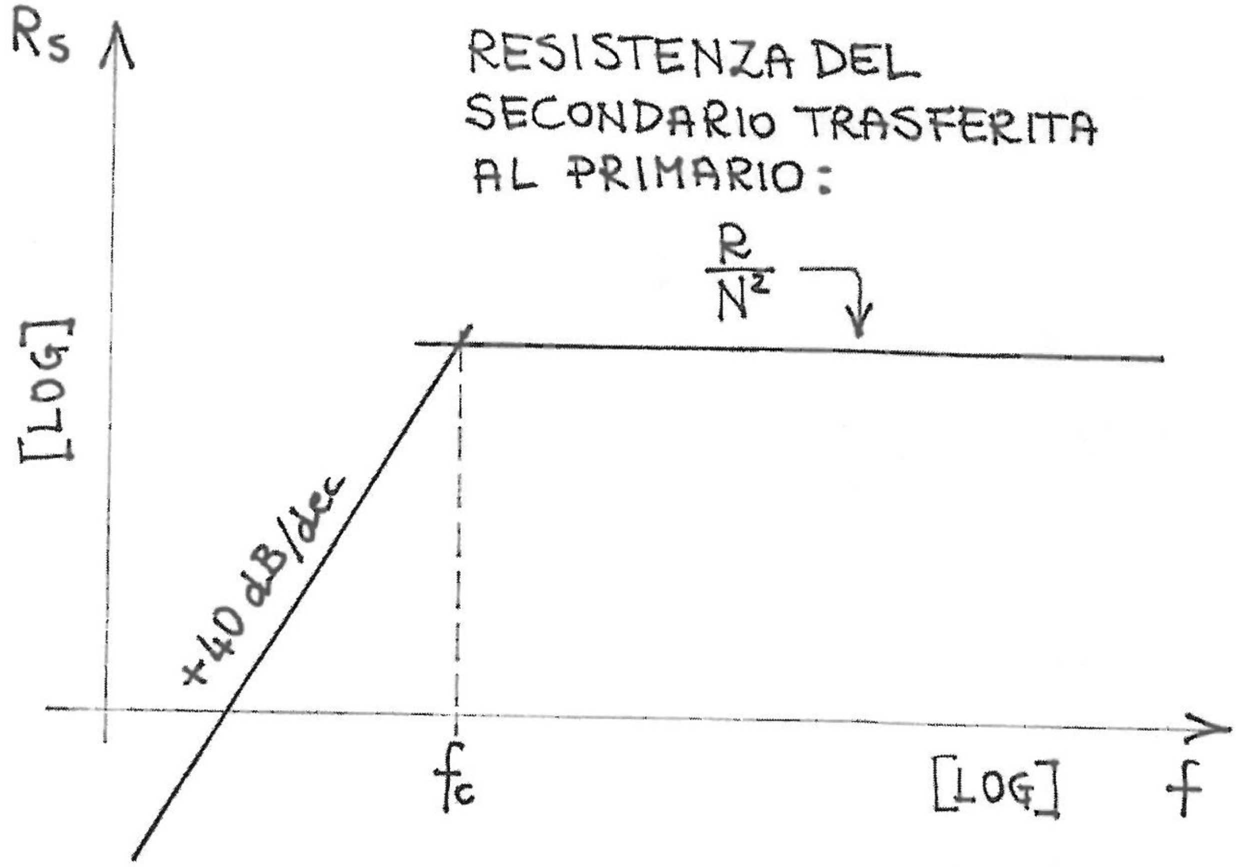
Residenza fisica



$$\boxed{\frac{V_V}{I} = j\omega L_p + Z_{ins} = Z_s}$$

(resistenza della spira parassita, R_p, assunta trascurabile)

Caratterizzazione analitica

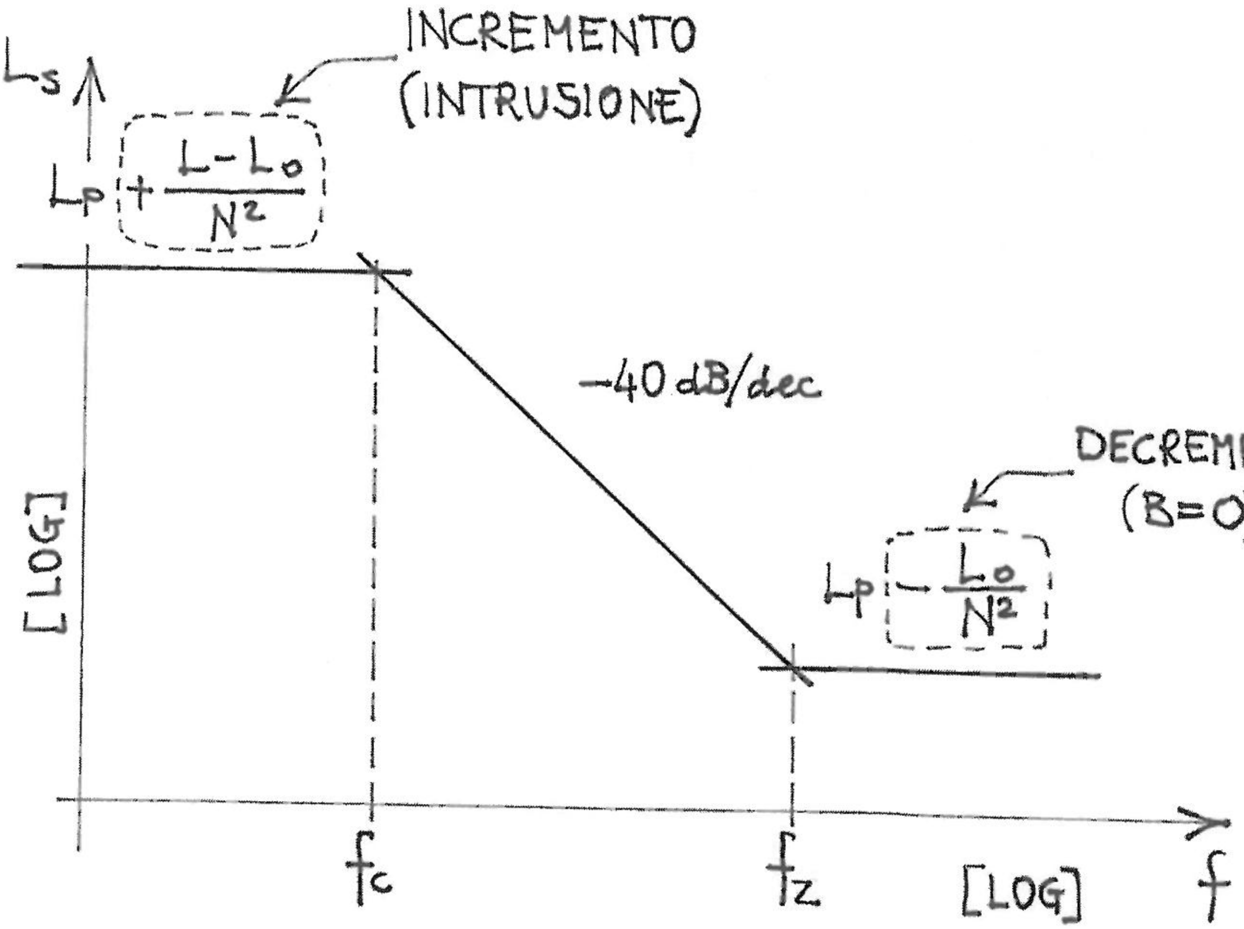


$$R_s = \frac{f^2}{f^2 + f_c^2} \cdot \frac{R}{N^2}, \qquad f_c = \frac{R}{2\pi L}$$

R_s È UN PARAMETRO EQUIVALENTE

<u>DIPENDENTE</u> <u>DALLA</u> <u>FREQUENZA</u>

Caratterizzazione analitica



$$L_s = \frac{f^2 + f_z^2}{f^2 + f_c^2} \cdot \left(L_p - \frac{L_o}{N^2} \right), \qquad f_z = f_c \cdot \sqrt{\frac{L_p + (L - L_o)/N^2}{L_p - L_o/N^2}} \gg f_c$$

L_s È UN PARAMETRO EQUIVALENTE

<u>DIPENDENTE</u> <u>DALLA</u> <u>FREQUENZA</u>

Parassiti - Impedenza di corto-circuito (*hp* 4193A)

Confronto caratterizzazione analitica - misura

Per fare il confronto occorre conoscere i seguenti ingredienti:

Resistenza di richiusura del secondario della sonda R

Numero di spire N

Induttanza del secondario L

Valore asintotico dell'induttanza parassita $L_s^{HF} = L_p - L_o/N^2$

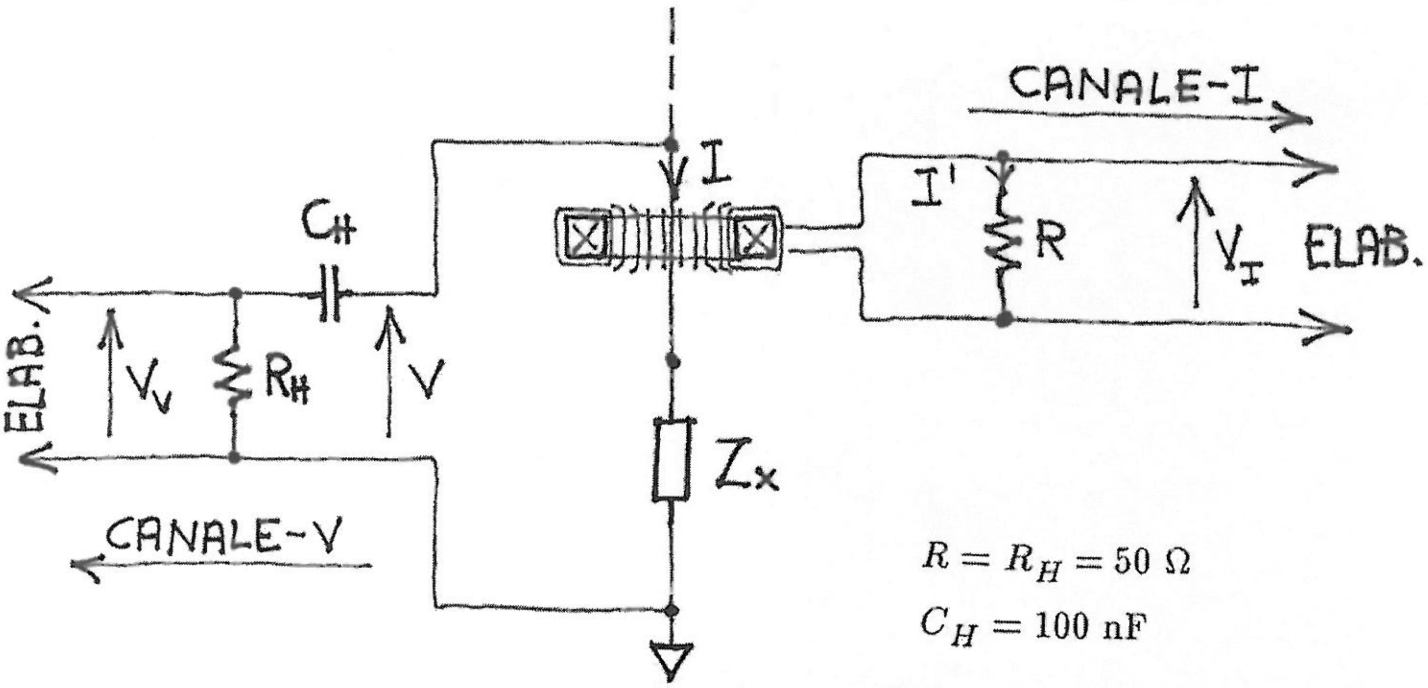
- Dal manuale di servizio dello strumento:

$$R = 50\ \Omega$$

- Dalla misura di Z_s si ricavano i valori asintotici:

$$R_s^{HF} = R/N^2 = 0.5\ \Omega,\ \text{da cui}\ N = 10$$
$$L_s^{HF} = 8.3\ \text{nH}$$

<u>Rimane da determinare L, autoinduttanza del secondario</u>



$R = R_H = 50\ \Omega$

$C_H = 100$ nF

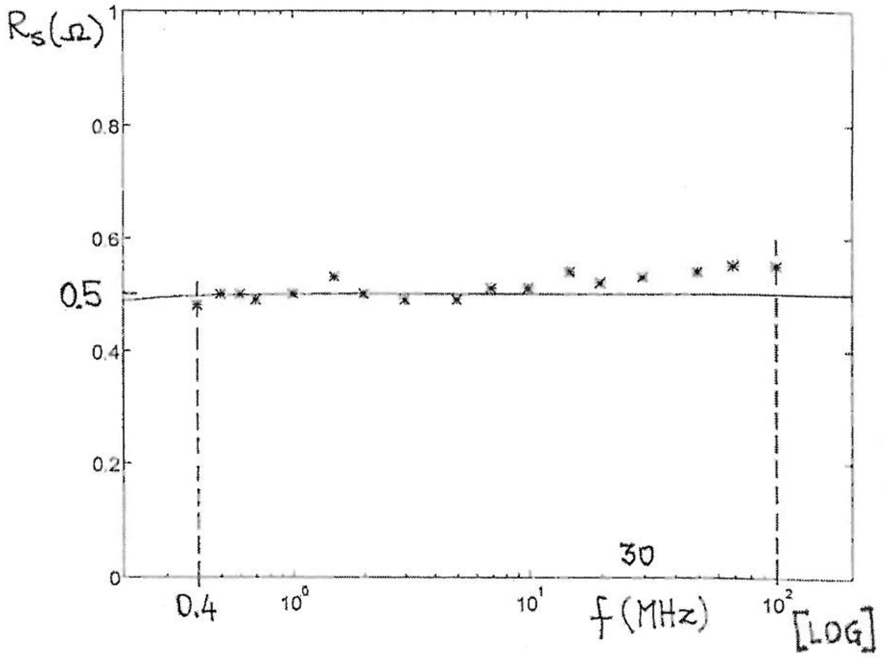
IL FILTRO PASSA ALTO RIPRODUCE NEL CANALE DI TENSIONE L'EFFETTO PASSA ALTO DELLA SONDA DI CORRENTE

$$\boxed{\frac{V_V}{V} = \frac{V_I}{I} \cdot \frac{1}{\frac{R}{N}} = \frac{Z_T}{\frac{R}{N}}}$$

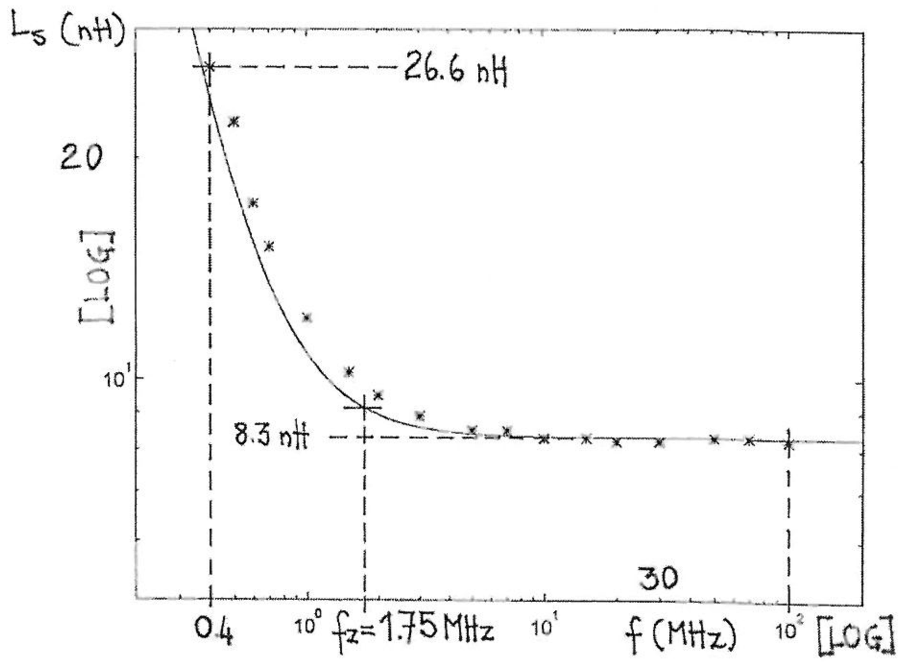
$$f_c = \frac{1}{2\pi R_H C_H} = \frac{R}{2\pi L} = 31.9 \text{ kHz, da cui } L = 250\ \mu\text{H}$$

Parassiti - Impedenza di corto-circuito (*hp* 4193A)

Confronto caratterizzazione analitica - misura



Scarto medio calcolo-misura = 14 mΩ



Scarto medio calcolo-misura = 0.76 nH

Conclusioni

CAPACITÀ PARASSITA DI CIRCUITO APERTO:

a) modella l'effetto delle correnti di spostamento fra la superficie calda e fredda della sonda (ed eventuali adattatori) dell'impedenzimetro

b) dipende dalle caratteristiche geometriche della sonda e degli adattatori

c) non dipende dalla frequenza (perché è una capacità)

Effetti parassiti nelle misure di impedenza vettoriali

Conclusioni

IMPEDENZA DI CORTO-CIRCUITO, COMPONENTE RESISTIVA:

a) è la parte reale dell'impedenza di inserzione della sonda di corrente

b) è costante nell'intervallo di frequenze di misura dell'impedenzimetro *hp* 4193A (400 kHz $\div$ 110 MHz)

IMPEDENZA DI CORTO-CIRCUITO, COMPONENTE INDUTTIVA:

a) è la somma dell'induttanza della spira parassita (primario) con l'induttanza apparente dell'impedenza di inserzione

b) nell'intervallo di frequenze di misura dell'impedenzimetro cala al crescere della frequenza tendendo asintoticamente ad una valore costante

Table I-1. Specifications. (Sheet 1 of 5)

SPECIFICATIONS

IMPEDANCE MAGNITUDE MEASUREMENT:

Range, Display, and Resolution:

MAGNITUDE RANGE	DISPLAY RANGE	DISPLAY (digit)	RESOLUTION
10Ω	00.00Ω to 19.99Ω	3 1/2	10mΩ
100Ω	000.0Ω to 199.9Ω	3 1/2	100mΩ
1kΩ	0.000kΩ to 1.999kΩ	3 1/2	1Ω
10kΩ	00.00kΩ to 19.99kΩ	3 1/2	10Ω
100kΩ	000.kΩ to 120.kΩ	2 1/2	1kΩ

Accuracy: See Table A.

Range Mode: Auto and manual (up-down).

IMPEDANCE PHASE MEASUREMENT:

Range and Resolution:

MAGNITUDE RANGE	DISPLAY RANGE	RESOLUTION
10Ω	180.0° to -180.0°	0.1°
100Ω	180.0° to -180.0°	0.1°
1kΩ	180.0° to -180.0°	0.1°
10kΩ	180.0° to -180.0°	0.1°
100kΩ	180.° to -180.°	1°

Accuracy: See Table A.

Table A. Accuracies

MAGNITUDE RANGE		Test Frequency (MHz) 0.4 to 1	1 to 10	10 to 40	40 to 110
10Ω	Z	$\pm[(5.7+\frac{0.56}{f})$% of reading +9 counts]	±[6.3% of reading +6 counts]	±[(4.5+0.18f)% of reading +4 counts]	±[(4.5+0.18)% of reading +4 counts]
	θ	$\pm(1.7+\frac{1.8}{f}+\frac{35}{Z})$ degrees	$\pm(3.3+0.20f+\frac{35}{Z})$ degrees	$\pm(3.3+0.20f+\frac{35}{Z})$ degrees	$\pm(3.3+0.20f+\frac{35}{Z})$ degrees
100Ω	Z	$\pm[(2.4+\frac{0.56}{f})$% of reading +4 counts]	±[3.0% of reading +4 counts]	±[(2.6+0.037f)% of reading +4 counts]	±[(2.6+0.037f)% of reading +4 counts]
	θ	$\pm(1.5+\frac{1.9}{f}+\frac{35}{Z})$ degrees	$\pm(3.3+0.035f+\frac{35}{Z})$ degrees	$\pm(3.3+0.035f+\frac{35}{Z})$ degrees	$\pm(3.3+0.035f+\frac{35}{Z})$ degrees
1kΩ	Z	$\pm[(3.2+\frac{0.56}{f})$% of reading +4 counts]	±[3.7% of reading +4 counts]	±[(2.7+0.11f)% of reading +4 counts]	±[(2.7+0.11f)% of reading +4 counts]
	θ	$\pm(1.6+\frac{1.8}{f}+\frac{35}{Z})$ degrees	$\pm(3.3+0.11f+\frac{35}{Z})$ degrees	$\pm(3.3+0.11f+\frac{35}{Z})$ degrees	$\pm(3.3+0.11f+\frac{35}{Z})$ degrees
10kΩ	Z	$\pm[(2.9+\frac{0.56}{f})$% of reading +4 counts]	±[(3.2+0.29f)% of reading +4 counts]	±[(0.74+0.63f)% of reading +4 counts]	
	θ	$\pm(1.8+\frac{1.9}{f}+\frac{35}{Z})$ degrees	$\pm(3.1+0.53f+\frac{35}{Z})$ degrees	$\pm(8.3+0.01f+\frac{35}{Z})$ degrees	
100kΩ*	Z	$\pm[(3.3+\frac{0.56}{f})$% of reading +4 counts]			
	θ	$\pm(3.0+\frac{1.9}{f}+\frac{35}{Z})$ degrees			

Where, f is test frequency in MHz, and Z is number of MAGNITUDE display counts. On the 100kΩ range, the small zero "o" is not counted in Z.

*: Measurement accuracy is not specified above 100kΩ.

Table 1-2. General Information

SUPPLEMENTAL PERFORMANCE CHARACTERISTICS

MEASUREMENT TIME

Normal Mode : Approximately 1 sec. (typical)
High Speed Mode : Approximately 150 msec. (typical)

FREQUENCY SETTLING TIME

Approximately 5ms to 400ms

RANGING TIME

Approximately 1.2s

PROBE WITHSTAND VOLTAGE

DC : 50V maximum
AC : 5Vrms maximum

OUTPUT IMPEDANCE

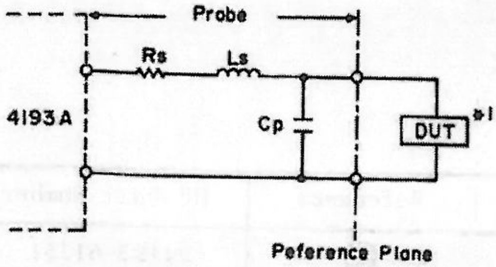
Approximately 25Ω with 0.2μF series capacitance

RESIDUALS

Resistance in series with DUT (Rs) : $\leq 0.55\Omega$

Inductance in series with DUT (Ls) : $\leq (4.9 + \frac{10}{f})$nH *2

Capacitance in parallel with DUT (Cp) : ≤ 0.1pF



Note
*1 : DUT includes the probe pin.
*2 : f is test frequency in MHz.

TEST SIGNAL LEVEL :

MAGNITUDE RANGE	CURRENT Thru DUT (μArms)
10Ω	100
100Ω	100
1kΩ	100
10kΩ	50
100kΩ	10

Note : Current through the DUT is constant for each magnitude range.

Accuracy : ±20%

RESIDUAL FM

100Hz_{p-p} for 1 thru 110MHz at 100Hz BW.

SKIP ERROR

10 counts maximum at 2.5MHz, 5MHz, and 10MHz.

Mutuo accoppiamento fra linee

Carlo Carobbi, Dicembre 2009

Si considera il caso di mutuo accoppiamento fra linee parallele e identiche, immerse in un dielettrico omogeneo e prive di perdite. La velocità di propagazione delle onde sulle linee è v. Le linee hanno lunghezza $\mathcal{L}$. I parametri per unità di lunghezza delle linee sono l'autoinduttanza l, la mutua induttanza l_m, la capacità c e la mutua capacità c_m. Si pone:

$$\mathbf{L} = \begin{bmatrix} l & l_m \\ l_m & l \end{bmatrix} \tag{1}$$

e

$$\mathbf{C} = \begin{bmatrix} c + c_m & -c_m \\ -c_m & c + c_m \end{bmatrix} \tag{2}$$

Le matrici $\mathbf{L}$ e $\mathbf{C}$ sono legate fra loro e si ha

$$\mathbf{LC} = \begin{bmatrix} 1/v^2 & 0 \\ 0 & 1/v^2 \end{bmatrix} \tag{3}$$

da cui, sostituendo la (1) e la (2) nella (3),

$$\begin{cases} l\left(c + c_m\right) - l_m c_m = \dfrac{1}{v^2} \\ -lc_m + l_m\left(c + c_m\right) = 0 \end{cases} \tag{4}$$

Dalle (4) si ottiene

$$\begin{cases} v = \dfrac{1}{\sqrt{1+k}} \dfrac{1}{\sqrt{lc}} \\ \dfrac{l_m}{l} = \dfrac{c_m}{c + c_m} \end{cases} \tag{5}$$

La lunghezza d'onda è $\lambda = v / f$, dove f è la frequenza, ed il numero d'onda è $\beta = 2\pi / \lambda$. Si definisce poi il fattore di mutuo accoppiamento k che è dato da

$$k = \frac{l_m}{l} = \frac{c_m}{c + c_m} \tag{6}$$

Assumiamo che una delle due linee sia alimentata da un generatore di tensione (linea G) e l'altra no (linea R). La resistenza interna del generatore che alimenta la linea G e le resistenze che chiudono le rimanenti tre porte (la porta non alimentata della linea G e le due porte della linea R) siano pari a Z_0, dove

$$Z_0 = \sqrt{\frac{l}{c+c_m}} = vl\sqrt{1-k^2} \tag{7}$$

e rappresenta l'impedenza caratteristica di ciascuna linea (in presenza dell'altra).
Risolvendo le equazioni delle linee accoppiate per determinare le tensioni alle porte della linea R si trova:

$$V_{NE} = \frac{V_G}{4}\left[\frac{j\omega l_m \mathcal{L}}{Z_0} + j\omega c_m \mathcal{L} Z_0\right]\frac{\sin(\beta\mathcal{L})}{\beta\mathcal{L}} \tag{8}$$

e

$$V_{FE} = \frac{V_G}{4}\left[-\frac{j\omega l_m \mathcal{L}}{Z_0} + j\omega c_m \mathcal{L} Z_0\right]\frac{\sin(\beta\mathcal{L})}{\beta\mathcal{L}}\frac{1}{\left[\cos(\beta\mathcal{L}) + j\frac{\sin(\beta\mathcal{L})}{\sqrt{1-k^2}}\right]^2} \tag{9}$$

dove V_G è la tensione di circuito aperto impressa dal generatore, V_{NE} è la tensione alla porta prossima al generatore mentre V_{FE} è la tensione alla porta opposta. Dalla seconda delle (5) e dalla (7) abbiamo

$$\frac{l_m}{c_m} = Z_0^2 \tag{10}$$

Se si sostituisce quindi la (10) nella (9) abbiamo $V_{FE} = 0$, mentre sostituendo la (10) nella (8) si trova

$$V_{NE} = j\frac{V_G}{2}\sin(\beta\mathcal{L})\frac{k}{\sqrt{1-k^2}} \tag{11}$$