The Fundamental Theory of Low Noise Oscillators with Special Reference to Some Detailed Designs

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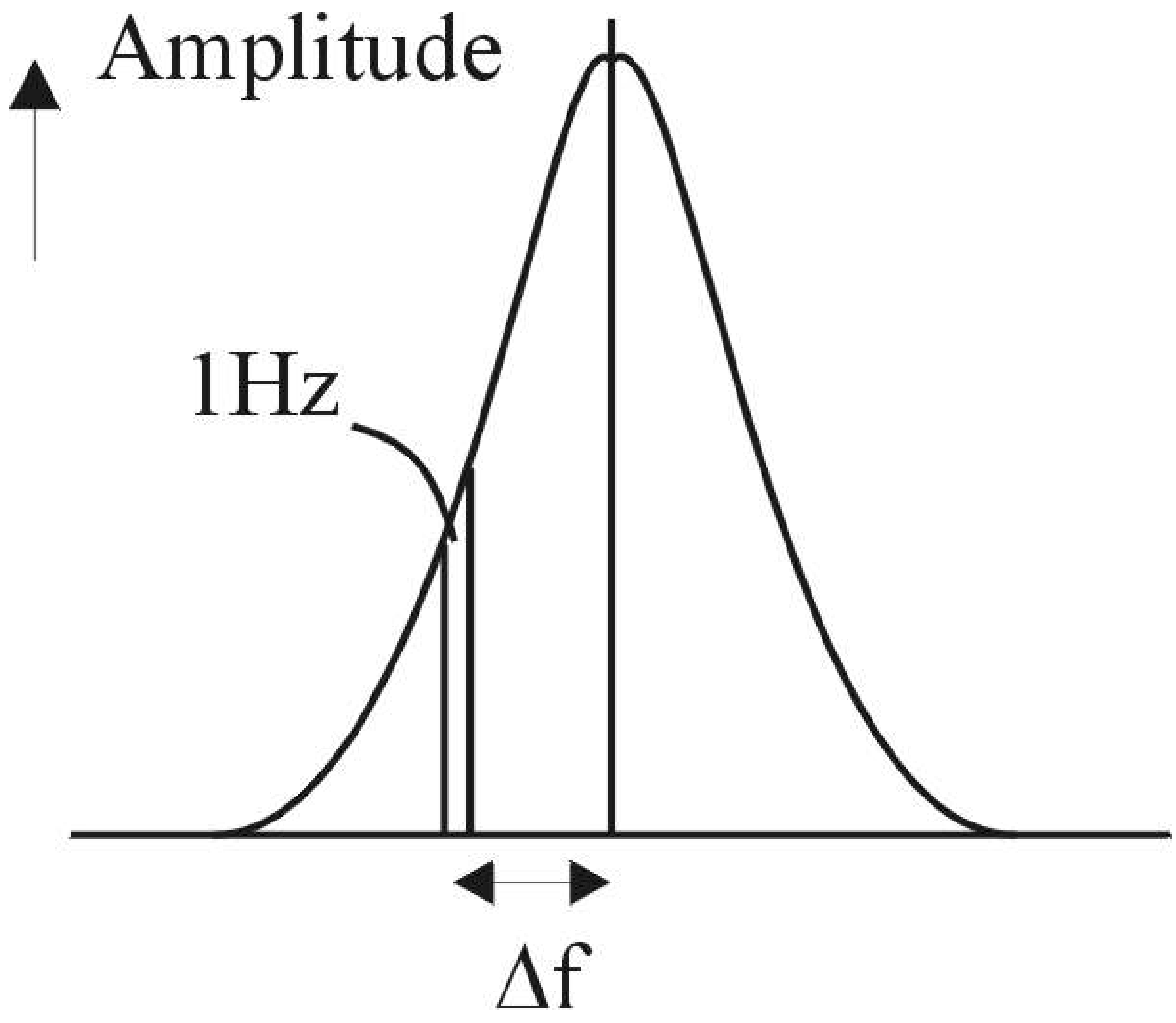
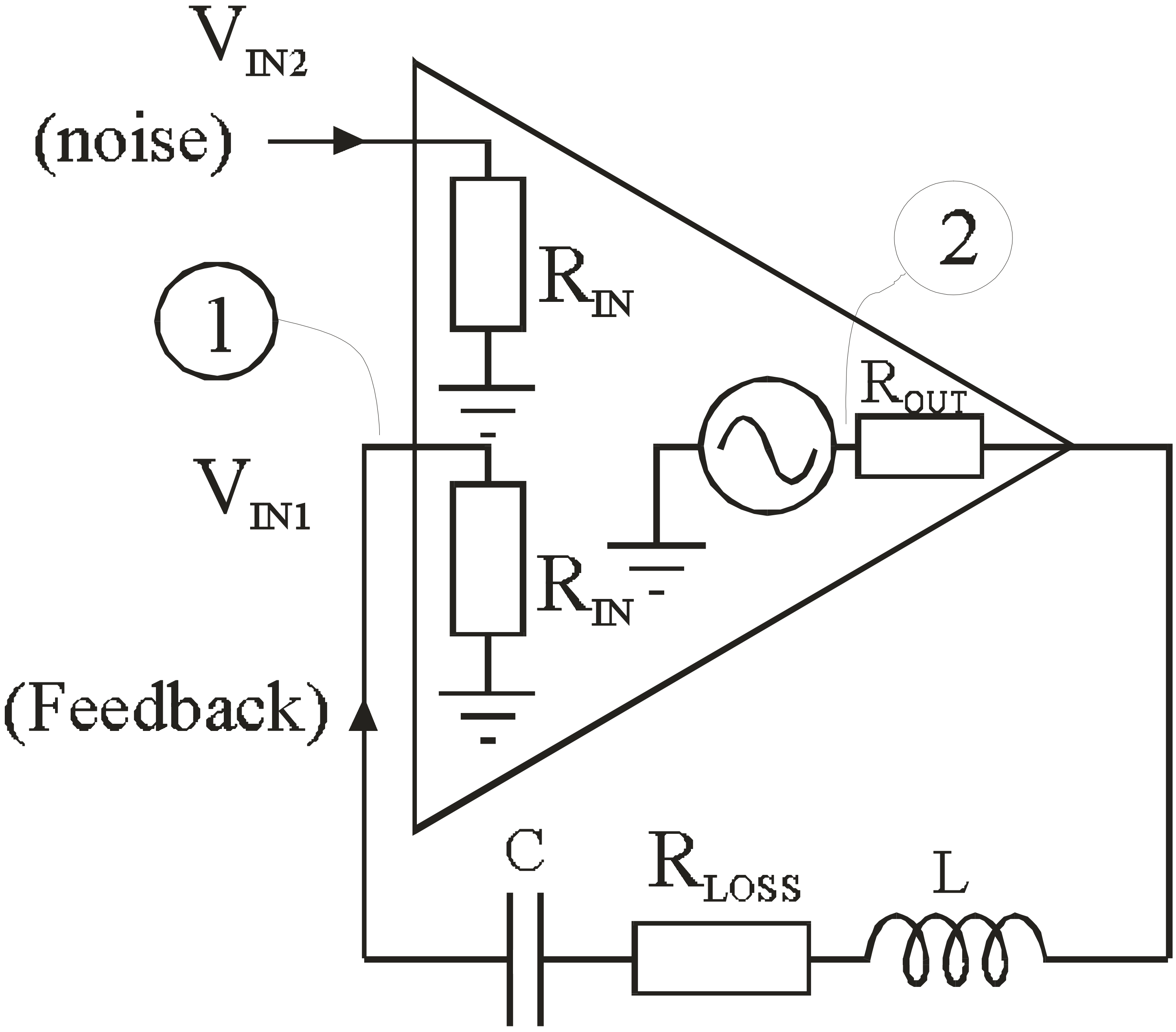
Low Noise Oscillators

- Oscillator models
- Noise theories for thermal (additive noise)
- Optimisation for minimum sideband noise
- Flicker noise measurement and reduction
- Oscillator designs
  - LC oscillators
  - SAW oscillators
  - Transmission line oscillators
- Tuning - varactor limitations
- Non-linear CAD
- Detailed designs

Spectrum of Oscillator

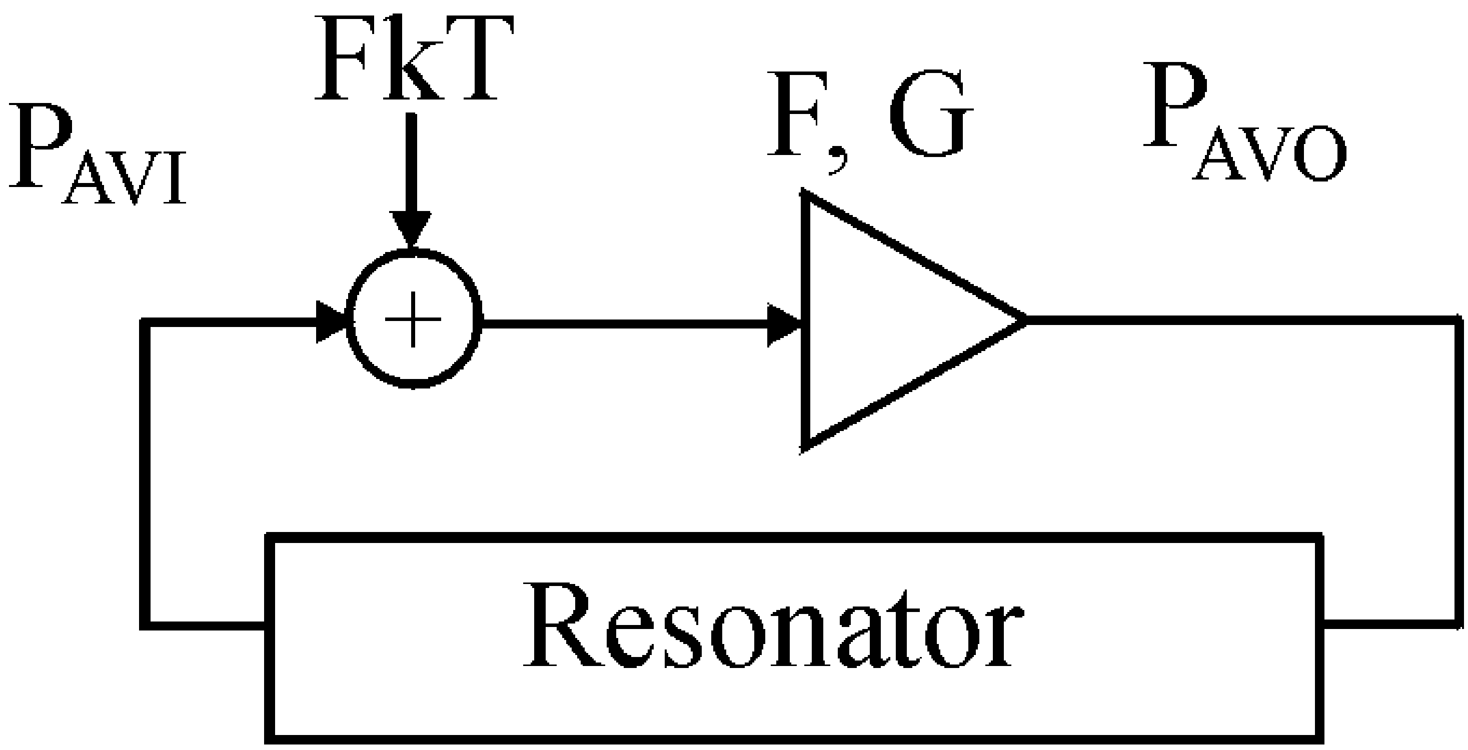


Oscillator Models

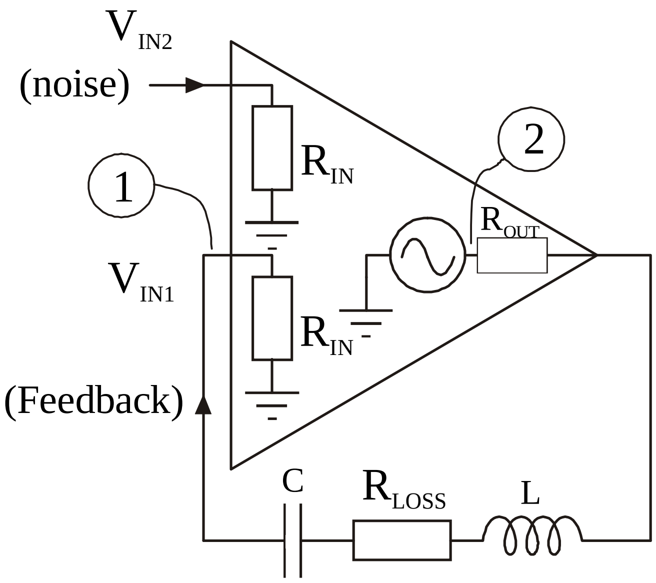


V_{IN2}
(noise)
R_{IN}
2
1
R_{OUT}
V_{IN1}
R_{IN}
(Feedback)
C
R_{LOSS}
L

Block model of oscillator



OSCILLATOR THEORY



$$\frac{V_{OUT}}{V_{IN2}} = \frac{G}{1-(\beta G)}$$

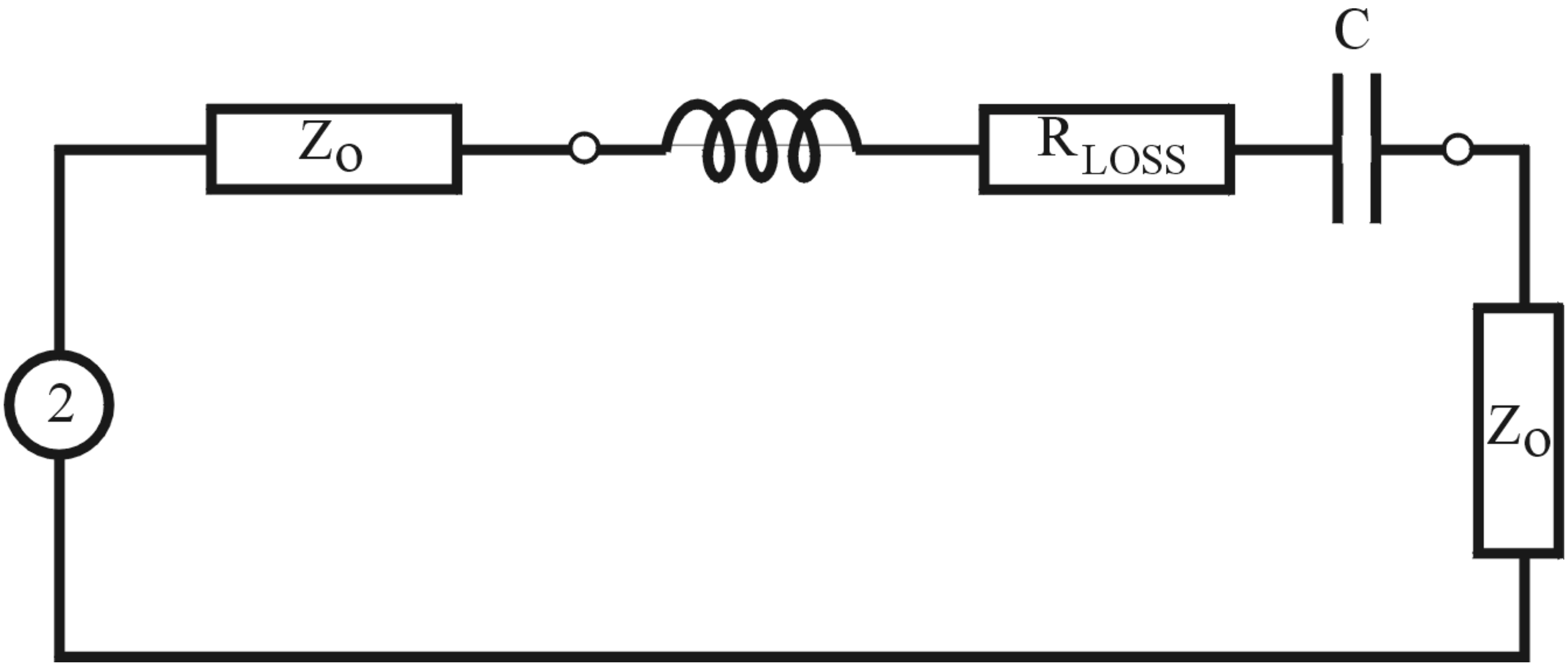
Model by Splitting original input into 2 identical inputs

One for noise injection

One for feedback

Model like Op-Amp with two inputs added and $\beta_0 G = 1$

Model of Resonator



Resonator Response

$$\beta = \left(\frac{R_{IN}}{R_{IN}+R_{OUT}}\right)\left(1-\frac{Q_L}{Q_0}\right)\frac{1}{\left(1 \pm 2jQ_L\dfrac{df}{f_o}\right)}$$

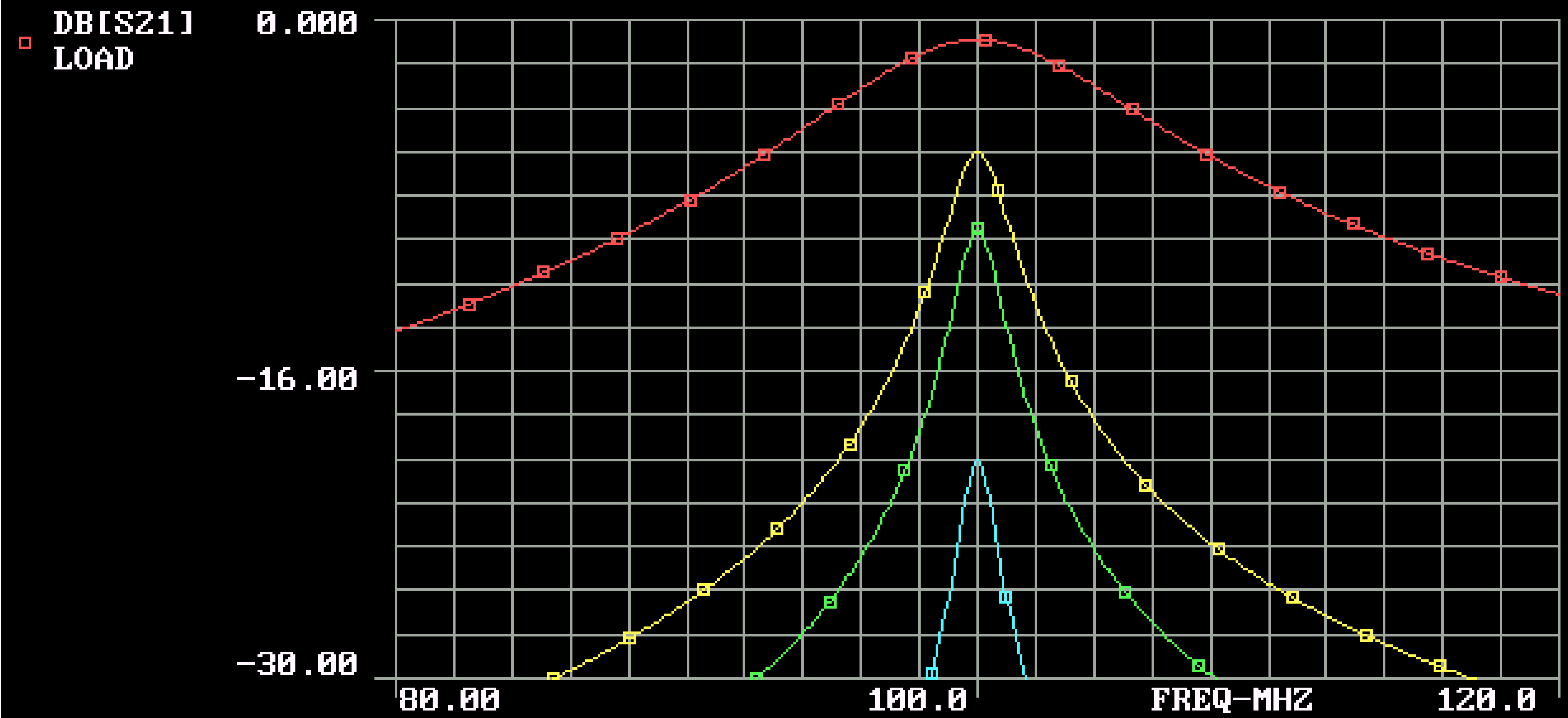
Insertion loss increases as Q_L tends to Q_0

If $R_{OUT} = R_{IN}$ then the insertion loss of the resonator is $S_{21} = 2\beta$, therefore:

$$S_{21} = \left(1-\frac{Q_L}{Q_0}\right)\frac{1}{\left(1 \pm 2jQ_L\dfrac{df}{f_o}\right)}$$

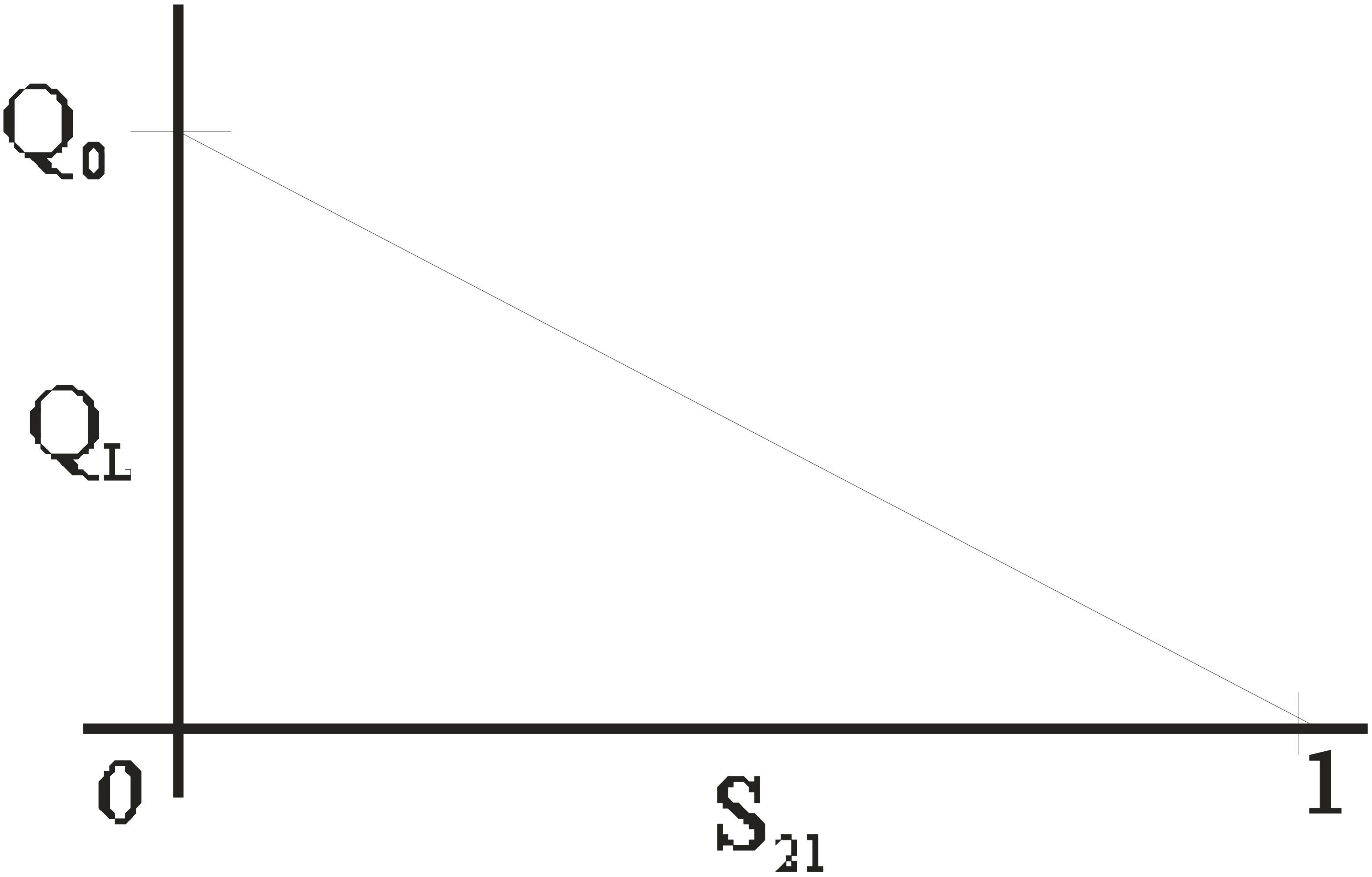
S_{21} = 6dB when Q_L/Q_0 = 1/2

S_{21} = 9dB when Q_L/Q_0 = 2/3



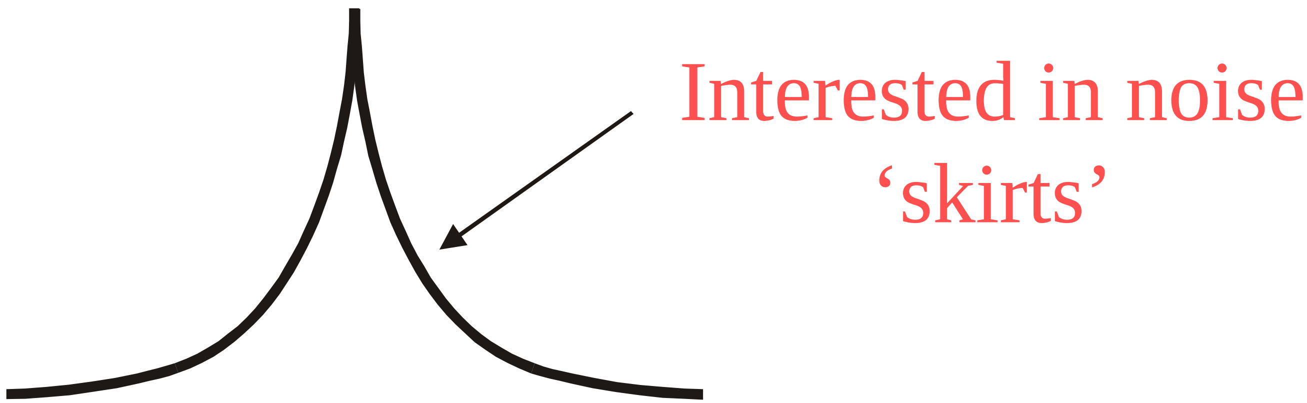
Resonator response versus Q_L/Q_0 = 0.1, 0.5, 2/3, 0.9

Insertion loss vs Q_L



At resonance Δf is zero and V_{OUT}/V_{IN2} is very large

$$\frac{V_{OUT}}{V_{IN2}} = \frac{G}{1 - \dfrac{1}{1 \pm \left(2jQ_L \dfrac{df}{f_0}\right)}} = \frac{1}{\left(1 - Q_L/Q_0\right)\left(\dfrac{R_{IN}}{R_{OUT}+R_{IN}}\right)\left(1 - \dfrac{1}{1 \pm \left(2jQ_L \dfrac{df}{f_0}\right)}\right)}$$



simplifies to:

$$\frac{V_{OUT}}{V_{IN2}} = \frac{G}{\pm 2jQ_L \dfrac{\Delta f}{f_o}} = \frac{1}{\left(1 - Q_L/Q_0\right)\left(\dfrac{R_{IN}}{R_{OUT}+R_{IN}}\right)\left(\pm 2jQ_L \dfrac{\Delta f}{f_0}\right)}$$

Noise in terms of power in 1Hz BW

- Calculate input noise power in 1Hz BW
  - initially calculate square of I/P voltage
- Assume O/P power limited
  - or at least always calculate in terms of O/P *
- Equation starts to break down very close to carrier at offsets typically << 1Hz
  - this is not usually a problem

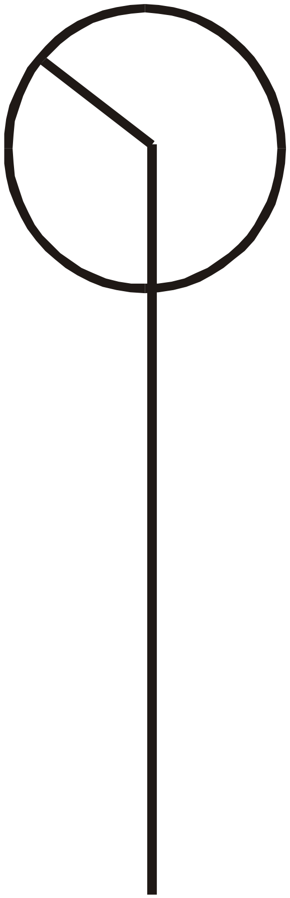
$$V_{IN} = \sqrt{FkTR_{IN}}$$

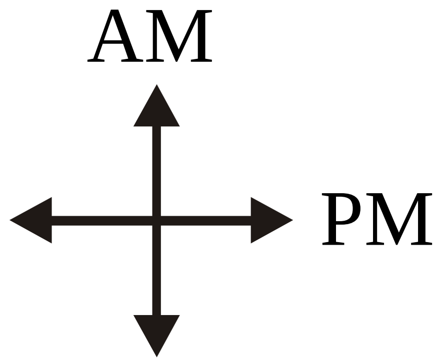
Noise input

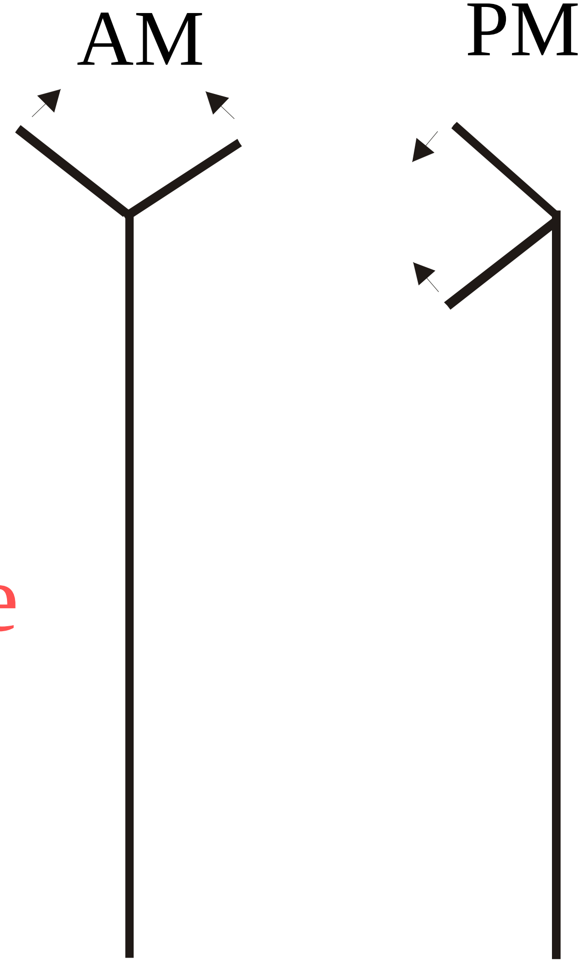
$$(V_{OUT}\,\Delta f)^2 = \frac{FkTR_{IN}}{4(Q_L)^2 (R_{IN}/(R_{OUT}+R_{IN}))^2 (1 - Q_L/Q_0)^2} \left(\frac{f_o}{\Delta f}\right)^2$$

Separate constants and variables

$$(V_{OUT}\,\Delta f)^2 = \frac{FkTR_{IN}}{4(Q_0)^2 (Q_L/Q_0)^2 (R_{IN}/(R_{OUT}+R_{IN}))^2 (1 - Q_L/Q_0)^2} \left(\frac{f_o}{\Delta f}\right)^2$$







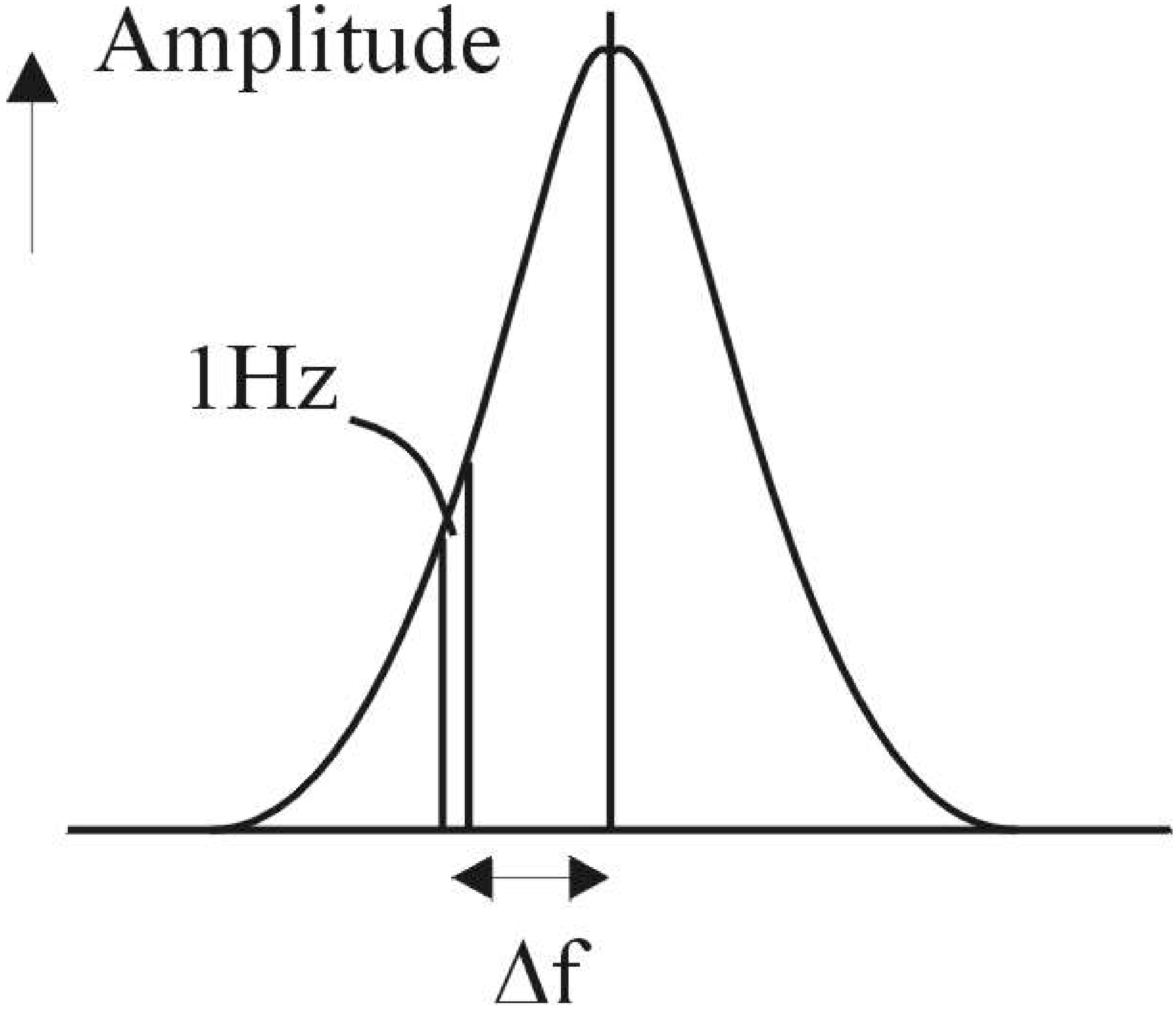
AM and PM noise model

$$(V_{OUT}\,\Delta f)^2 = \frac{FkTR_{IN}}{8(Q_0)^2\,(Q_L/Q_0)^2\,(R_{IN}/(R_{OUT}+R_{IN}))^2\,(1-Q_L/Q_0)^2}\left(\frac{f_o}{\Delta f}\right)^2$$

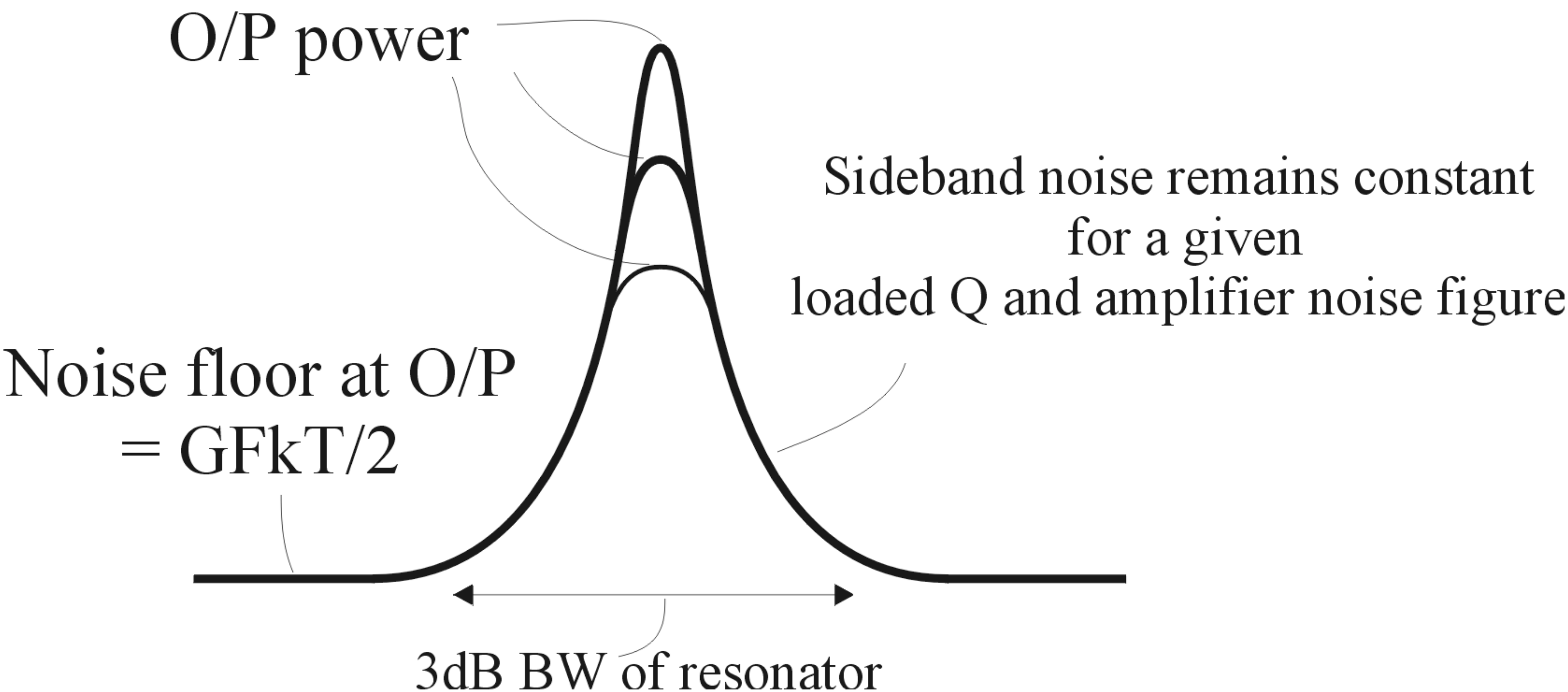
$$L_{FM} = \frac{\left(V_{OUT}\ \Delta f\right)^2}{\left(V_{OUT\ MAX\ RMS}\right)^2}$$

$$L_{FM} = \frac{FkTR_{IN}}{8\left(Q_0\right)^2\left(Q_L/Q_0\right)^2\left(1-Q_L/Q_0\right)^2\left(R_{IN}/\left(R_{OUT}+R_{IN}\right)\right)^2\left(V_{OUT\ MAX\ RMS}\right)^2}\left(\frac{f_o}{\Delta f}\right)^2$$

Define power P_{AVO} or P_{RF}


Amplitude
1Hz
Δf

Noise spectrum of oscillator



P_{RF}

$$P_{RF} = \frac{\left(V_{OUT\ MAX\ RMS}\right)^2}{R_{OUT} + R_{LOSS} + R_{IN}}$$

$$L_{FM} = \frac{FkT\left(R_{OUT} + R_{IN}\right)^2}{8(Q_0)^2\left(Q_L/Q_0\right)^2 R_{IN}\left(1 - Q_L/Q_0\right)^2 P_{RF}\left(R_{OUT} + R_{LOSS} + R_{IN}\right)}\left(\frac{f_O}{\Delta f}\right)^2$$

The ratio of sideband noise in a 1Hz BW at offset Δfto the total power is therefore:

R_{OUT} just dissipates power!

$$L_{FM} = \frac{FkT}{8(Q_0)^2\left(Q_L/Q_0\right)^2\left(1 - Q_L/Q_0\right)P_{RF}}\left(\frac{R_{OUT} + R_{IN}}{R_{IN}}\right)\left(\frac{f_O}{\Delta f}\right)^2$$

If R_{OUT} is zero as in a high efficiency oscillator

$$L_{FM} = \frac{FkT}{8(Q_0)^2 (Q_L/Q_0)^2 (1 - Q_L/Q_0) P_{RF}} \left(\frac{f_o}{\Delta f} \right)^2$$

If $R_{OUT} = R_{IN}$

Most amplifiers have similar I/P and O/P impedance

$$L_{FM} = \frac{FkT}{4(Q_0)^2 (Q_L/Q_0)^2 (1 - Q_L/Q_0) P_{RF}} \left(\frac{f_o}{\Delta f} \right)^2$$

Power available at the output P_{AVO} then:

$$P_{AVO} = \frac{\left(V_{OUT\ MAX\ RMS}\right)^2}{4\,R_{OUT}}$$

$$L_{FM} = \frac{FkT}{32(Q_0)^2\,(Q_L/Q_0)^2\,(1 - Q_L/Q_0)^2\,P_{AVO}}\left(\frac{(R_{OUT}+R_{IN})^2}{R_{OUT}\,.\,R_{IN}}\right)\left(\frac{f_o}{\Delta f}\right)^2$$

$$\left(\frac{(R_{OUT}+R_{IN})^2}{R_{OUT}\,.\,R_{IN}}\right) = 4$$

minimum when $R_{OUT} = R_{IN}$

If $R_{OUT} = R_{IN}$

$$L_{FM} = \frac{FkT}{8(Q_0)^2\,(Q_L/Q_0)^2\,(1 - Q_L/Q_0)^2\,P_{AVO}}\left(\frac{f_o}{\Delta f}\right)^2$$

General equation which describes all three cases

$$L_{FM} = A \cdot \frac{FkT}{8\left(Q_0\right)^2\left(Q_L/Q_0\right)^2\left(1 - Q_L/Q_0\right)^N P}\left(\frac{f_0}{\Delta f}\right)^2$$

1. N = 1 and A = 1 if P is defined as P_{RF} and R_{OUT} = zero
2. N = 1 and A = 2 if P is defined as P_{RF} and $R_{OUT} = R_{IN}$
3. N = 2 and A = 1 if P is defined as P_{AVO} and $R_{OUT} = R_{IN}$

The effect of the load

- Load not included so far
- Incorporate as coupler/attenuator at O/P of amplifier which causes:
  - Reduction in open loop gain
  - Increase in amplifier noise figure
  - NB Closed loop gain does not change as this is set by the insertion loss of the resonator
- Effect of load reduced if amplifier has zero/low O/P impedance

Optimisation for minimum noise

- The amplifier gain and resonator loaded Q are directly linked:

$$S_{21} = \left(1 - Q_L / Q_0\right)$$

- The noise factor is also dependent on loaded Q due to the change in source impedance
  - This is a second order effect and will be considered later

OPTIMISATION FOR MINIMUM NOISE

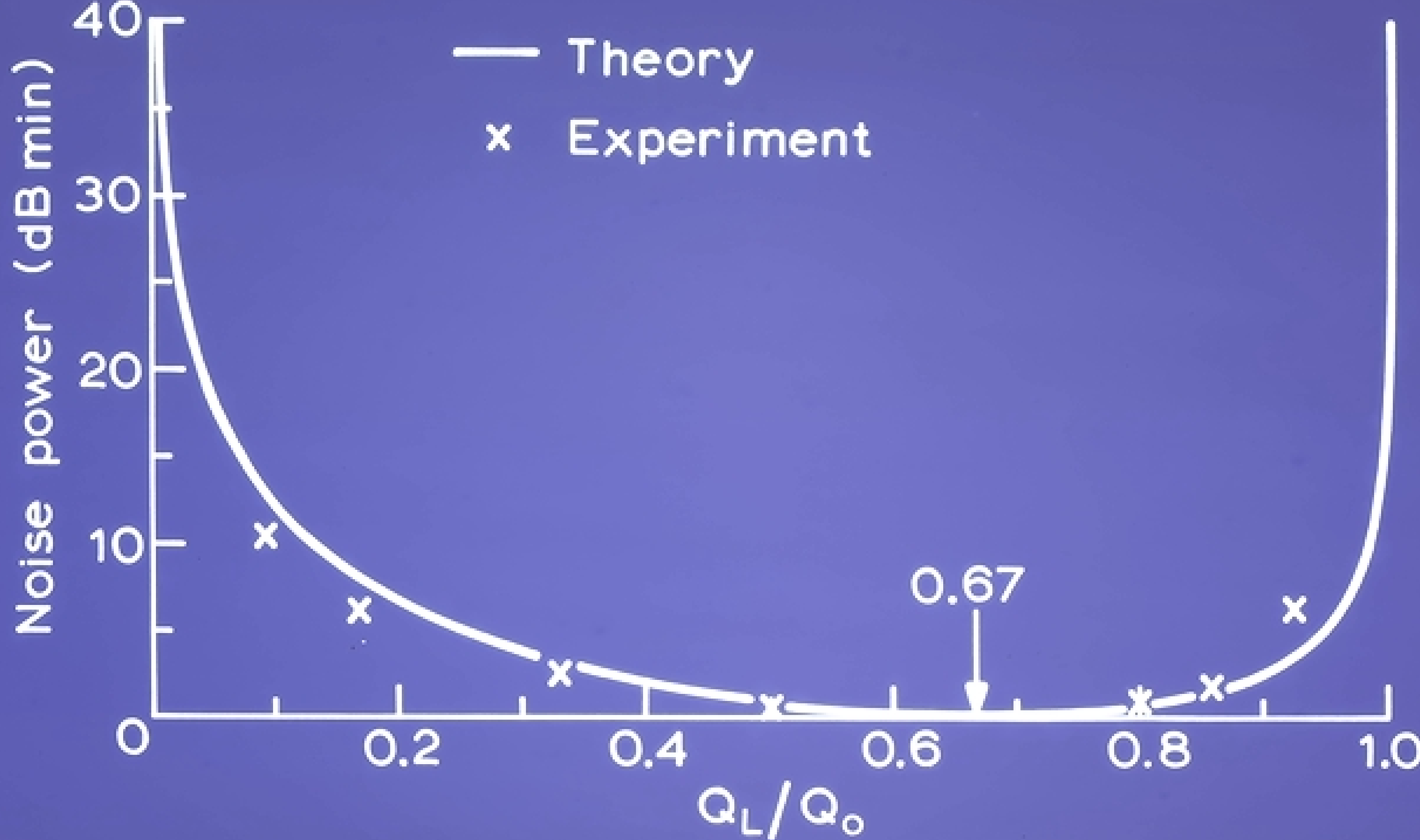
$$L_{(fm)} = \frac{FKT}{8Q_o^2 (Q_L / Q_o)^2 (1 - Q_L / Q_o) PFED} \times \left[\frac{f_o}{\delta f}\right]^2$$

For minimum noise if F is constant

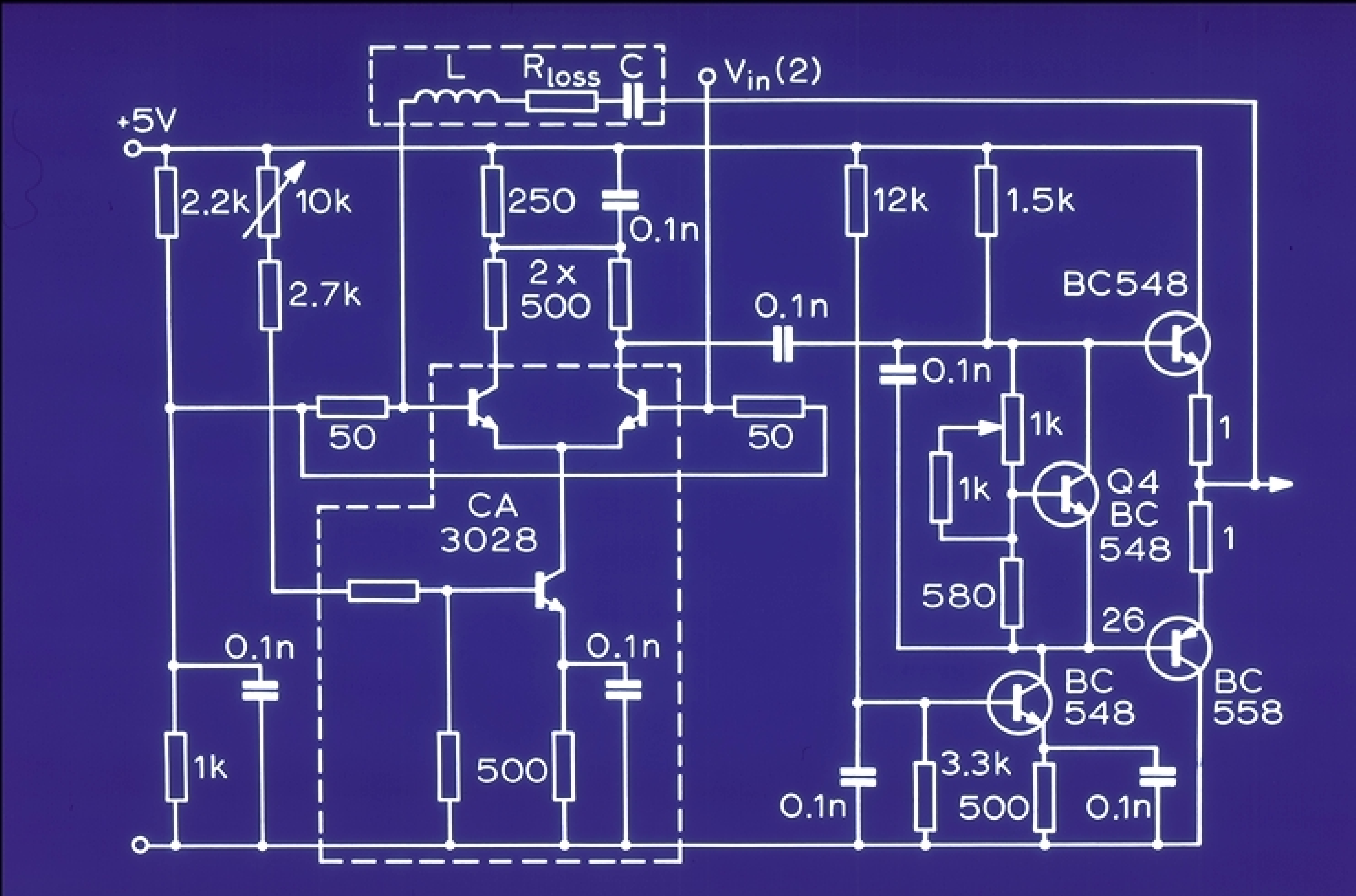
$$\frac{\delta (L_{(fm)})}{\delta (Q_L / Q_o)} = 0$$

MINIMUM NOISE OCCURS WHEN:

$Q_L / Q_o = 2/3$ and $G = 3$

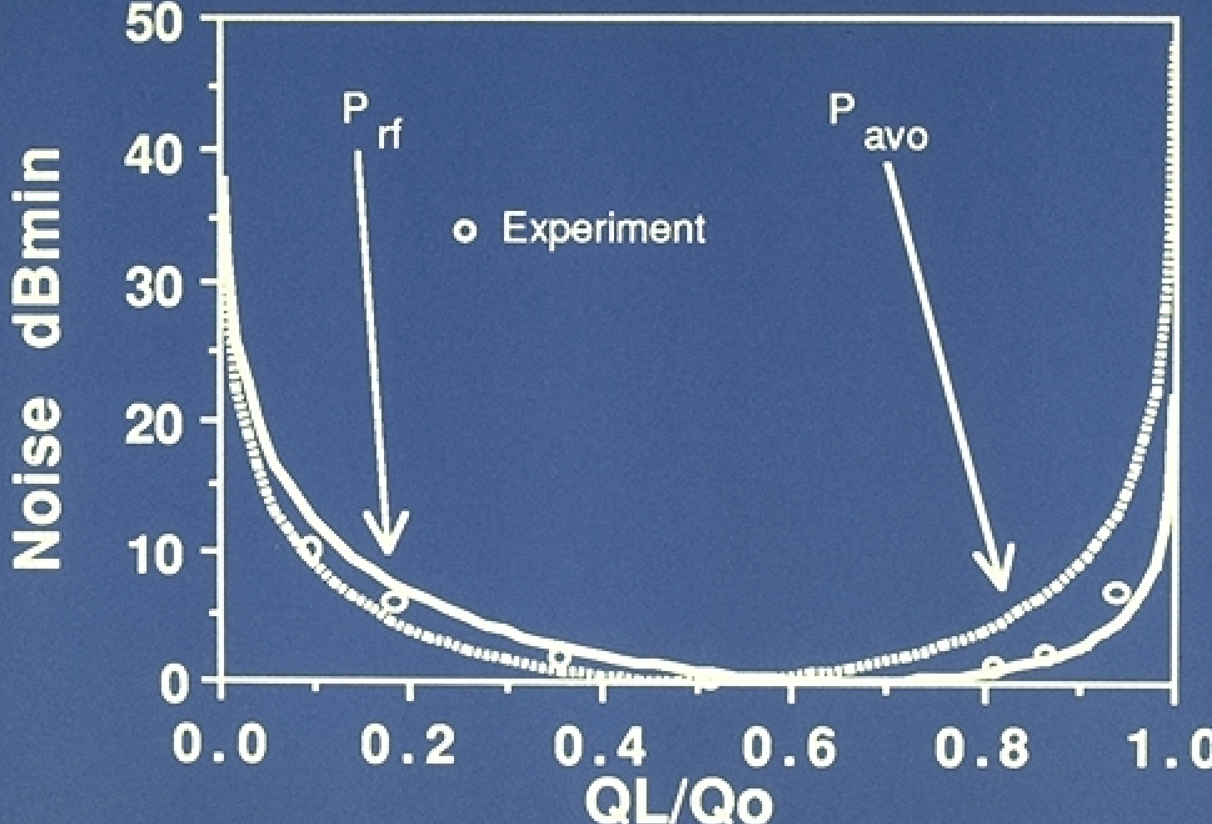


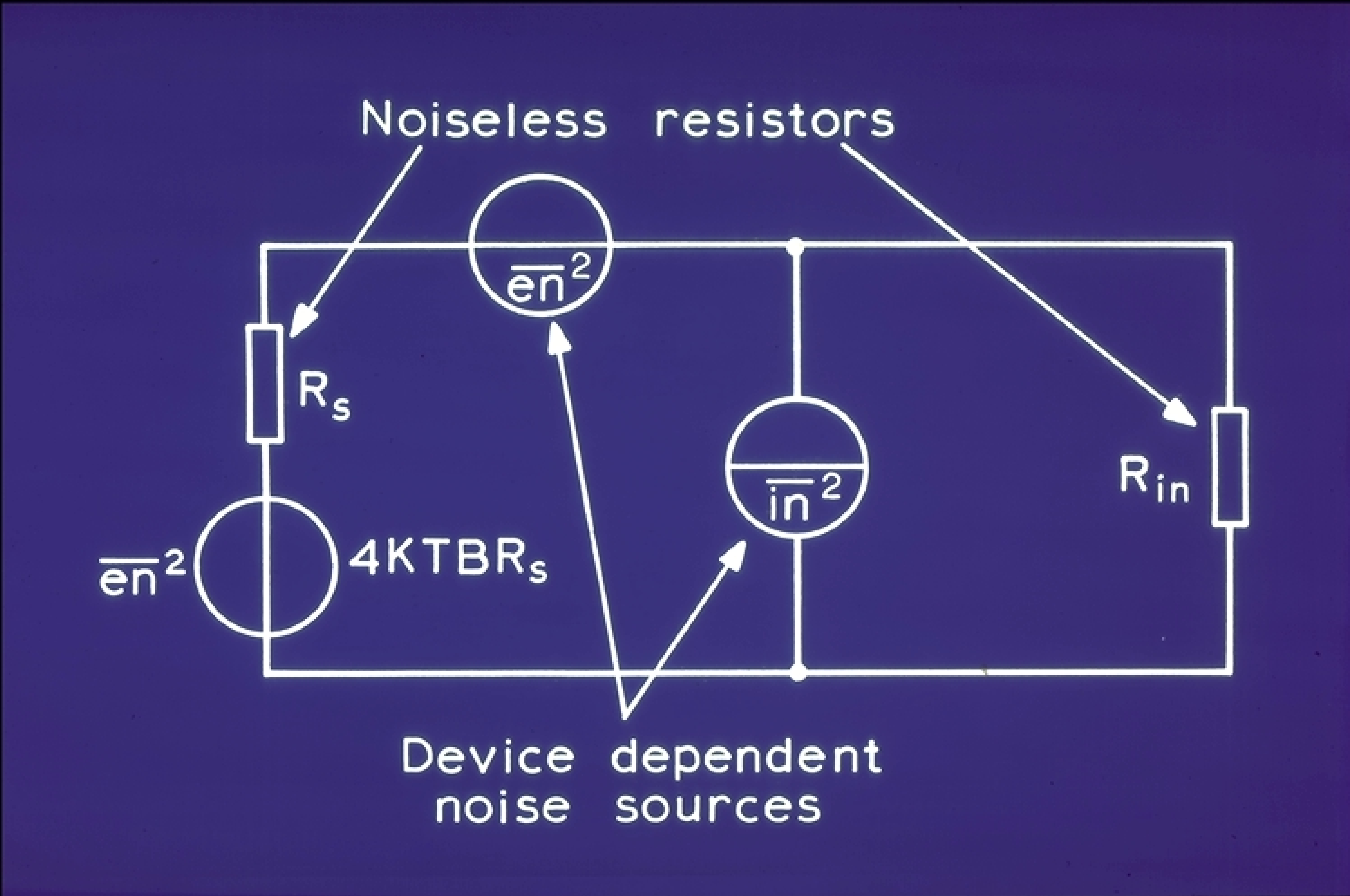
Sideband noise v. Q_L/Q_o



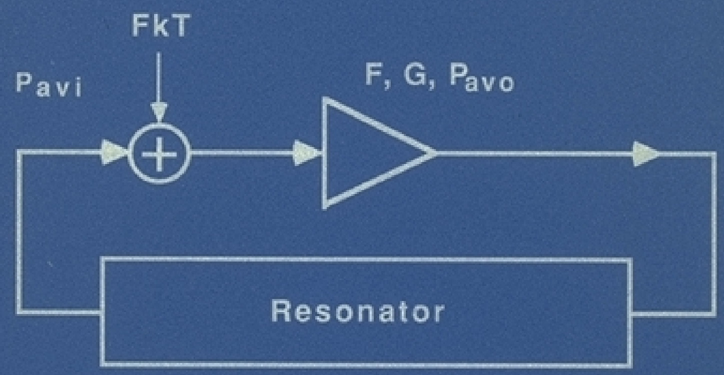
L
R$_{loss}$
C
V$_{in}$(2)
+5V
2.2k
10k
250
0.1n
12k
1.5k
2.7k
2 x
500
0.1n
BC548
0.1n
50
50
1k
1k
Q4
BC
548
1
1
CA
3028
580
26
0.1n
0.1n
BC
548
BC
558
1k
500
3.3k
500
0.1n
0.1n

Noise vs QL/Qo





Noiseless resistors
$\overline{en}^2$
R_s
$\overline{in}^2$
R_{in}
$\overline{en}^2$
$4KTBR_s$
Device dependent noise sources



What happens if the power that is limited is defined as Power available at the input of the amplifier

The noise equation now becomes

$$L_{fm} = \frac{FkT}{8Q_L^2 P_{avi}} (f_o/\delta f)^2$$

but the gain has now disappeared

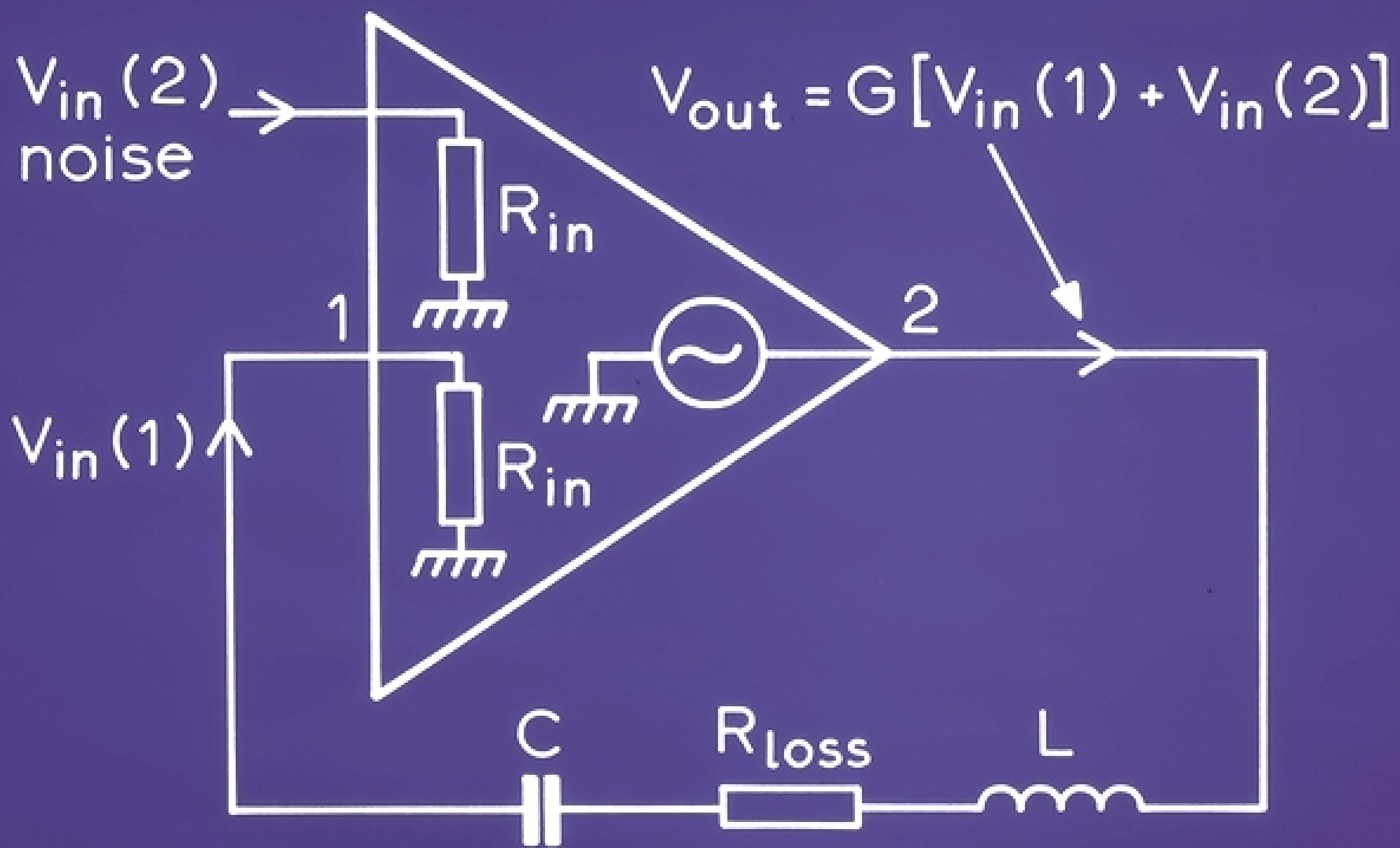
One therefore expects that Q_L should be high and made close to Q_o

However if Q_L tends to Q_o,

the amplifier gain and hence power have to be

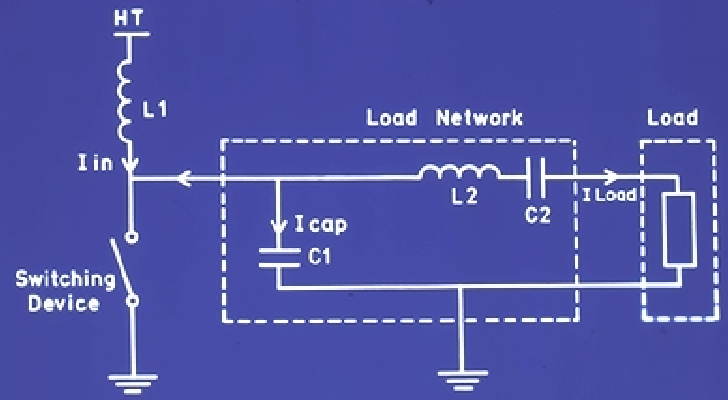
Infinite

High efficiency 1GHz oscillator

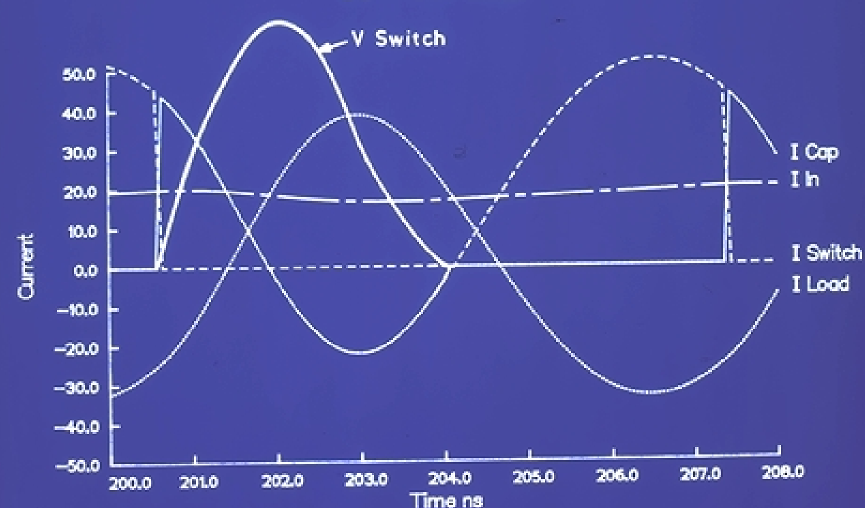


Oscillator model.

Basic Class E Amplifier



Class E Current Waveforms
Fc = 147 MHz



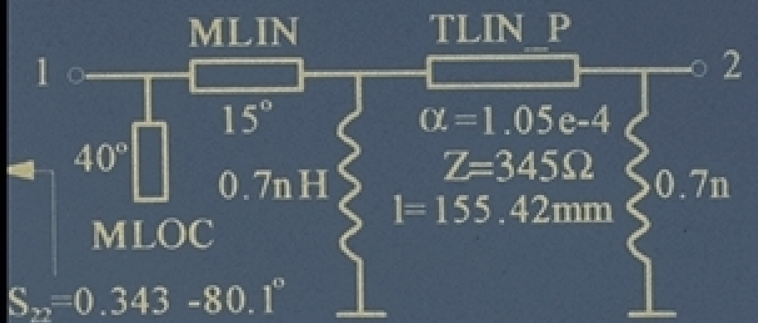


Figure 4 High Q Class E amplifier Load Network

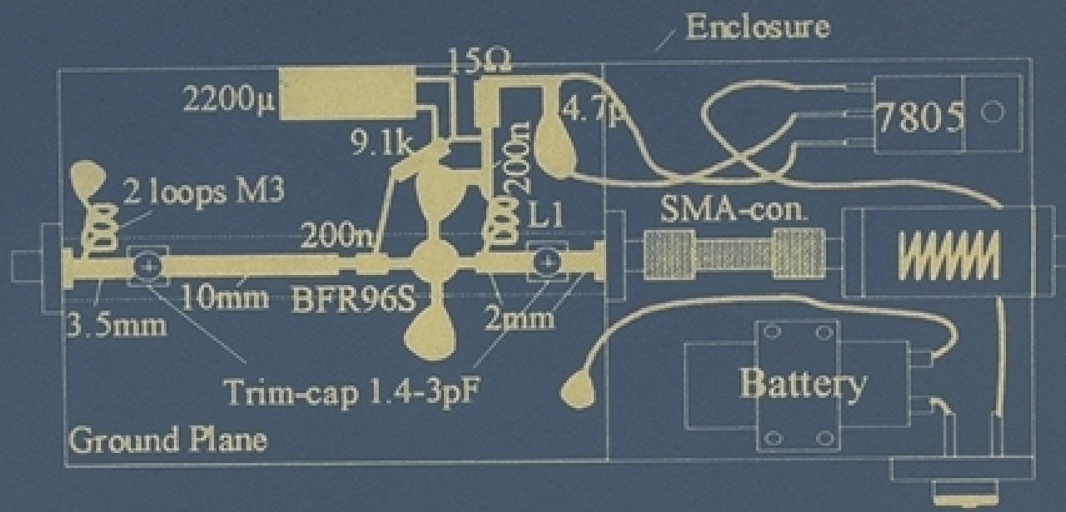
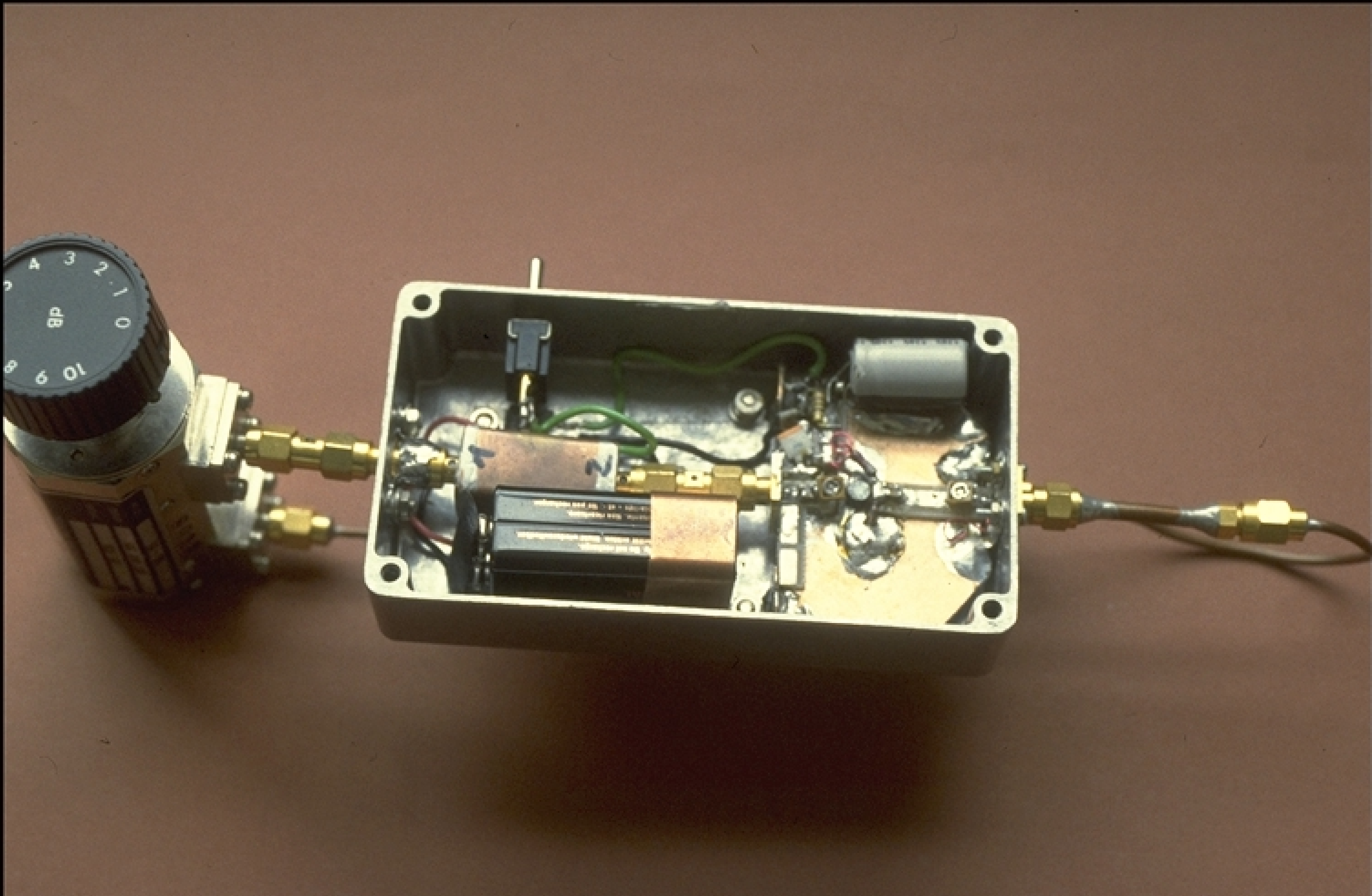


Figure 5, PCB layout


BP

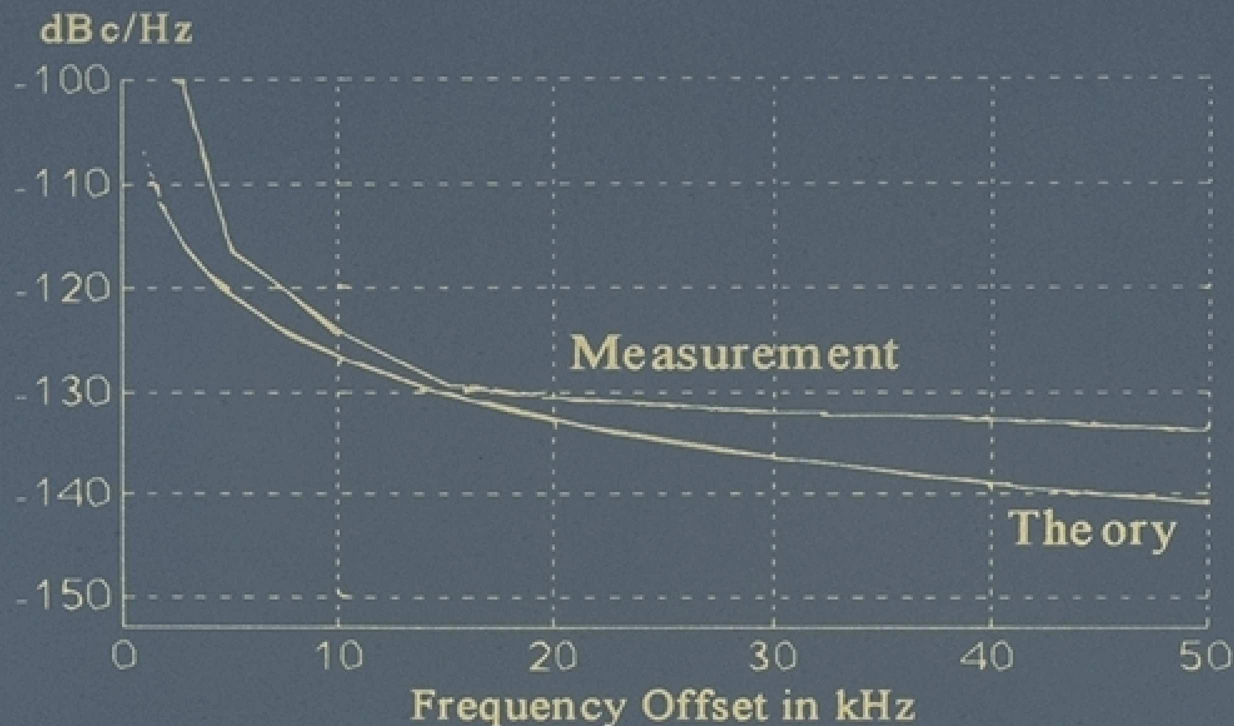


Figure 14 Phase Noise Performance of Optimised Oscillator

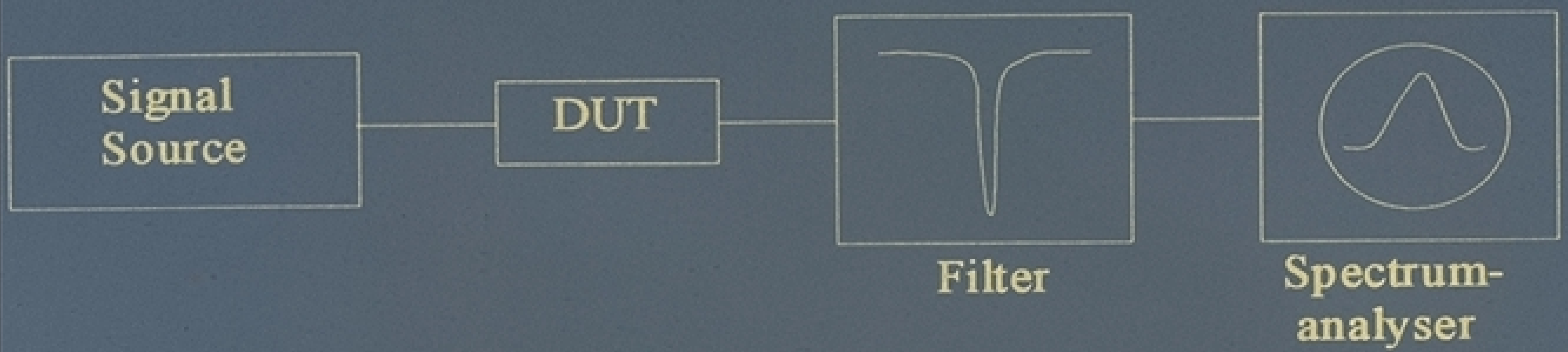
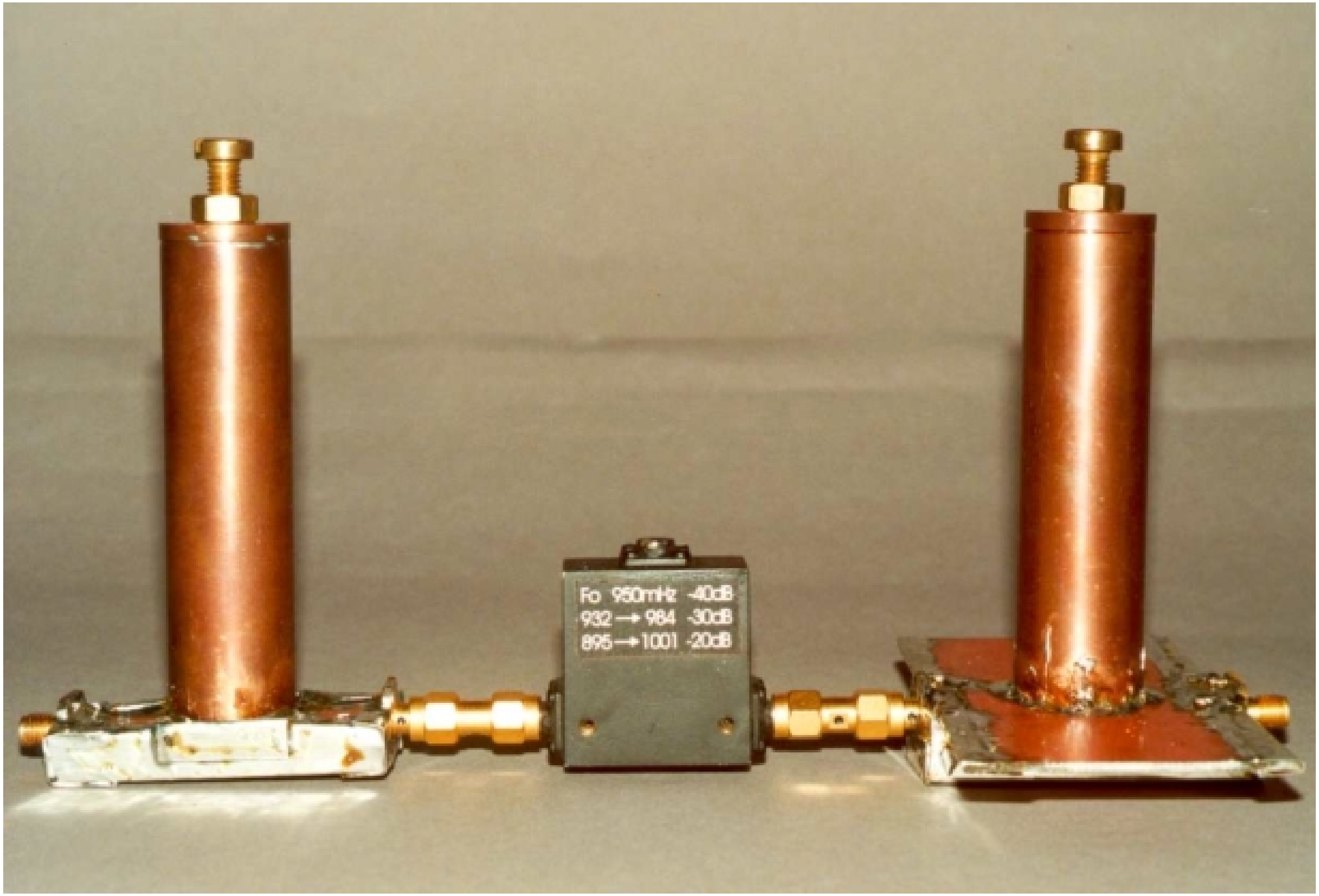
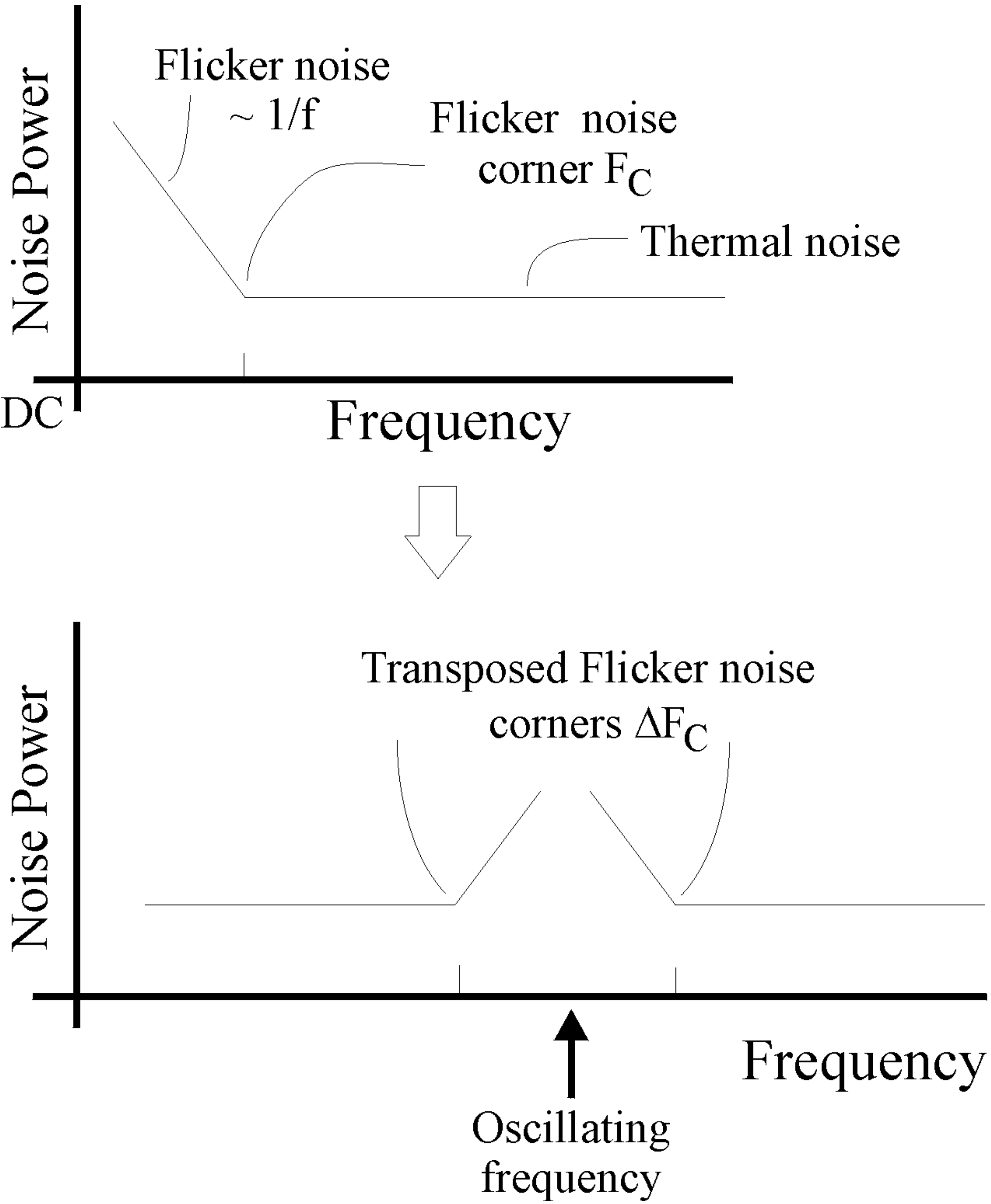


Figure 11 Noise Figure Measurement System.

Double transmission line filter



Flicker noise: measurement and reduction



Transposed flicker noise

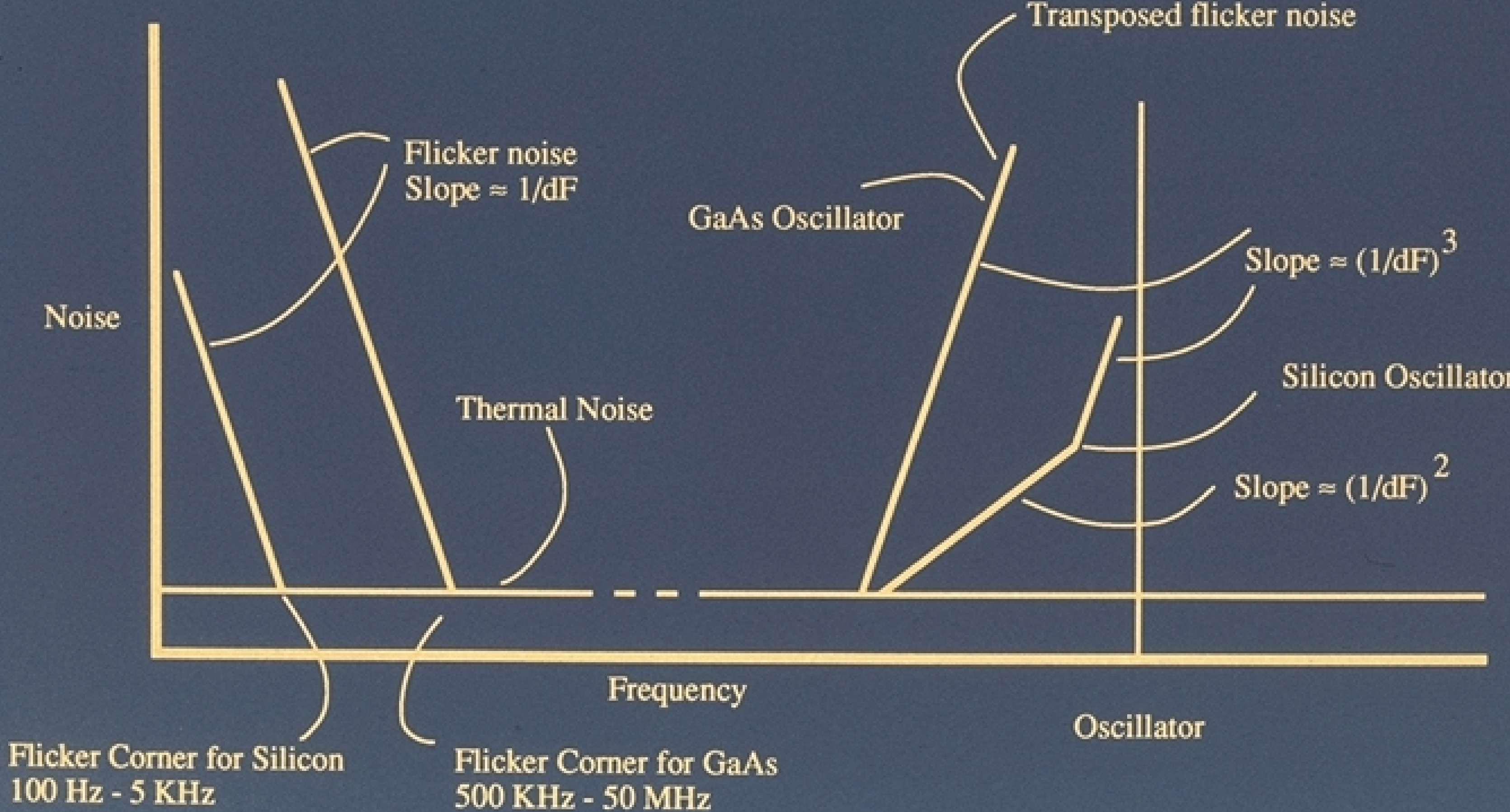
Low frequency flicker noise has ~1/f characteristic with flicker noise corner of F_C

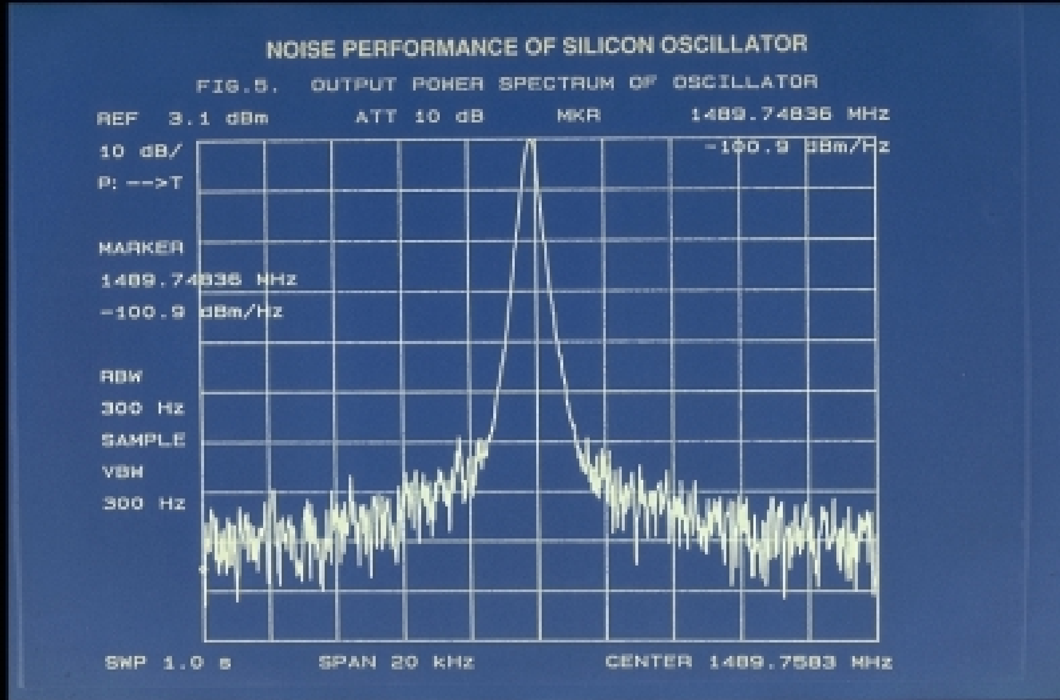
This is modulated onto carrier causing transposed flicker noise with 1/Δf characteristic

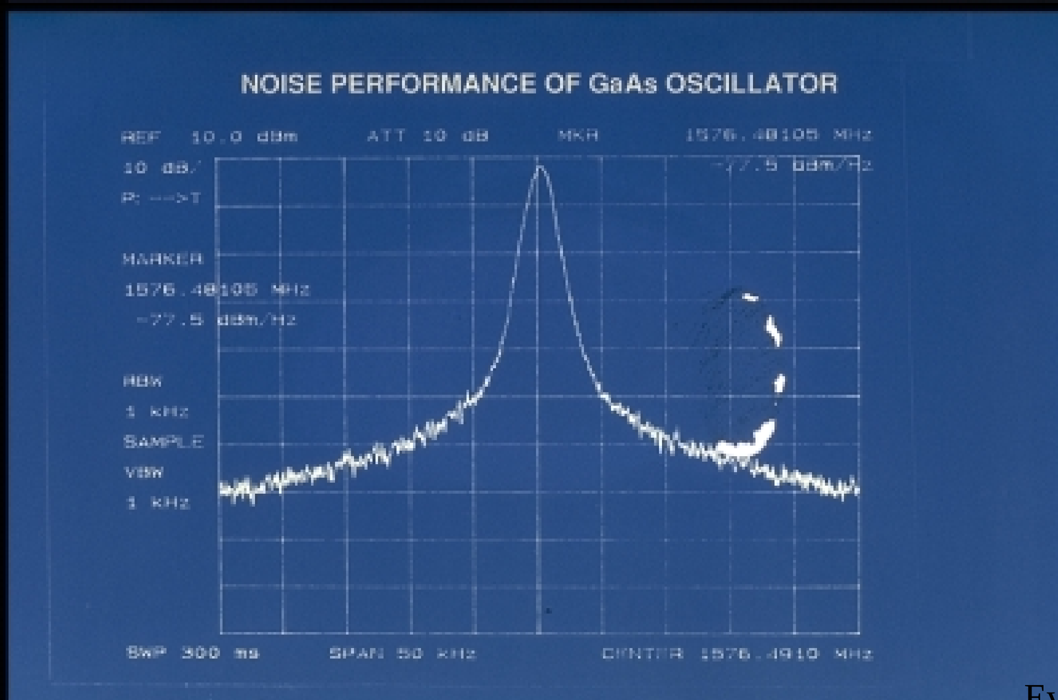
The transposed flicker noise corner ΔF_C is not the same as F_C

In oscillator this causes $(1/\Delta F)^3$ characteristic at offsets below ΔF_C

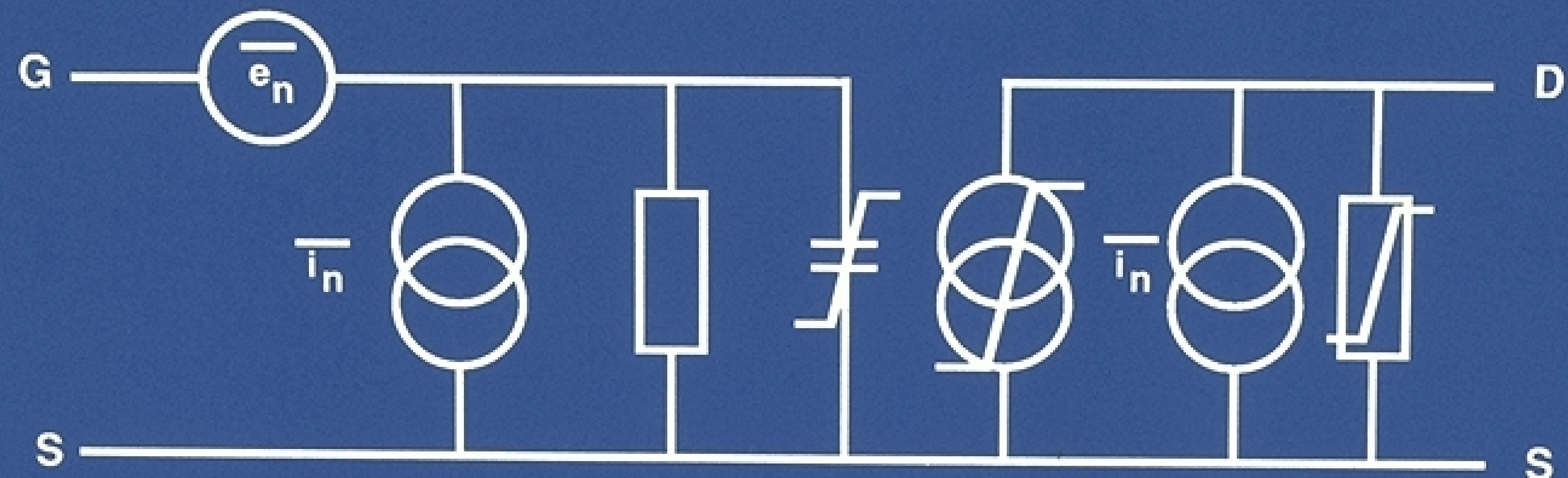
FLICKER NOISE PRODUCES NOISE DEGRADATION IN OSCILLATORS




NOISE PERFORMANCE OF SILICON OSCILLATOR
FIG.5. OUTPUT POWER SPECTRUM OF OSCILLATOR
REF 3.1 dBm
ATT 10 dB
MKR
1489.74836 MHz
-100.9 dBm/Hz
10 dB/
P: -->T
MARKER
1489.74836 MHz
-100.9 dBm/Hz
RBW
300 Hz
SAMPLE
VBW
300 Hz
SWP 1.0 s
SPAN 20 kHz
CENTER 1489.7583 MHz

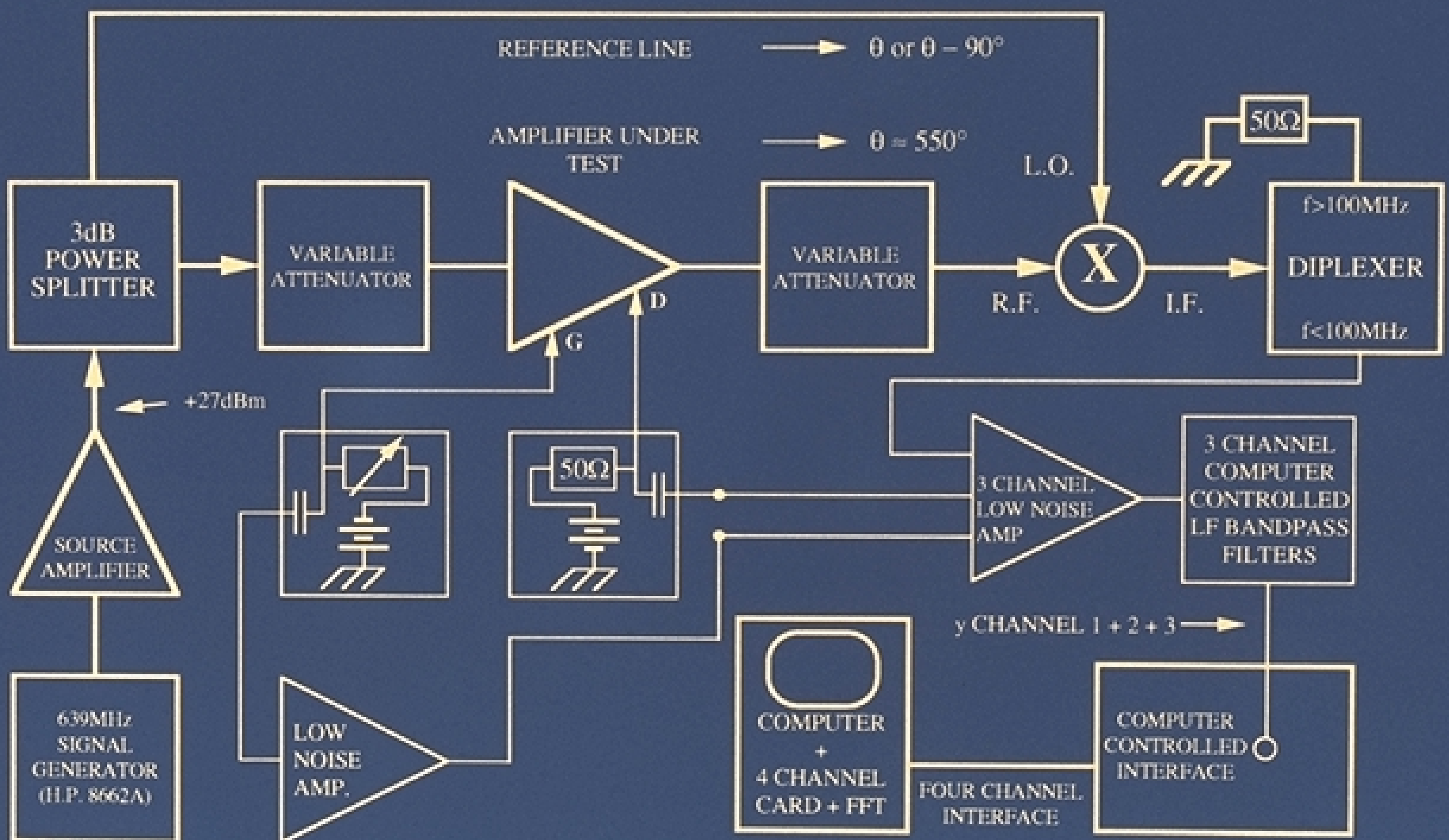

NOISE PERFORMANCE OF GaAs OSCILLATOR
REF 10.0 dBm
ATT 10 dB
MKR
1576.48105 MHz
-77.5 dBm/Hz
10 dB/
P: -->T
MARKER
1576.48105 MHz
-77.5 dBm/Hz
RBW
1 kHz
SAMPLE
VBW
1 kHz
SWP 300 ms
SPAN 50 kHz
CENTER 1576.4910 MHz

Eve

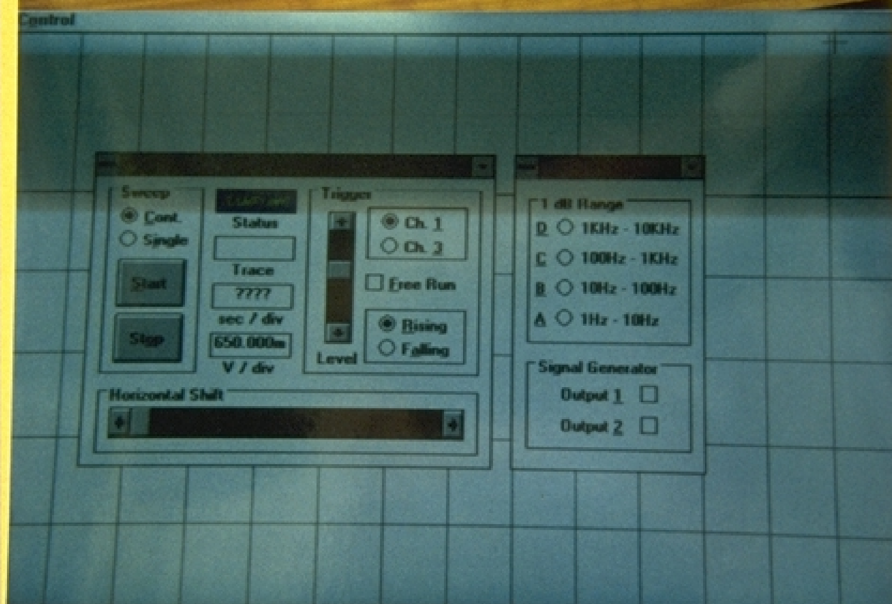


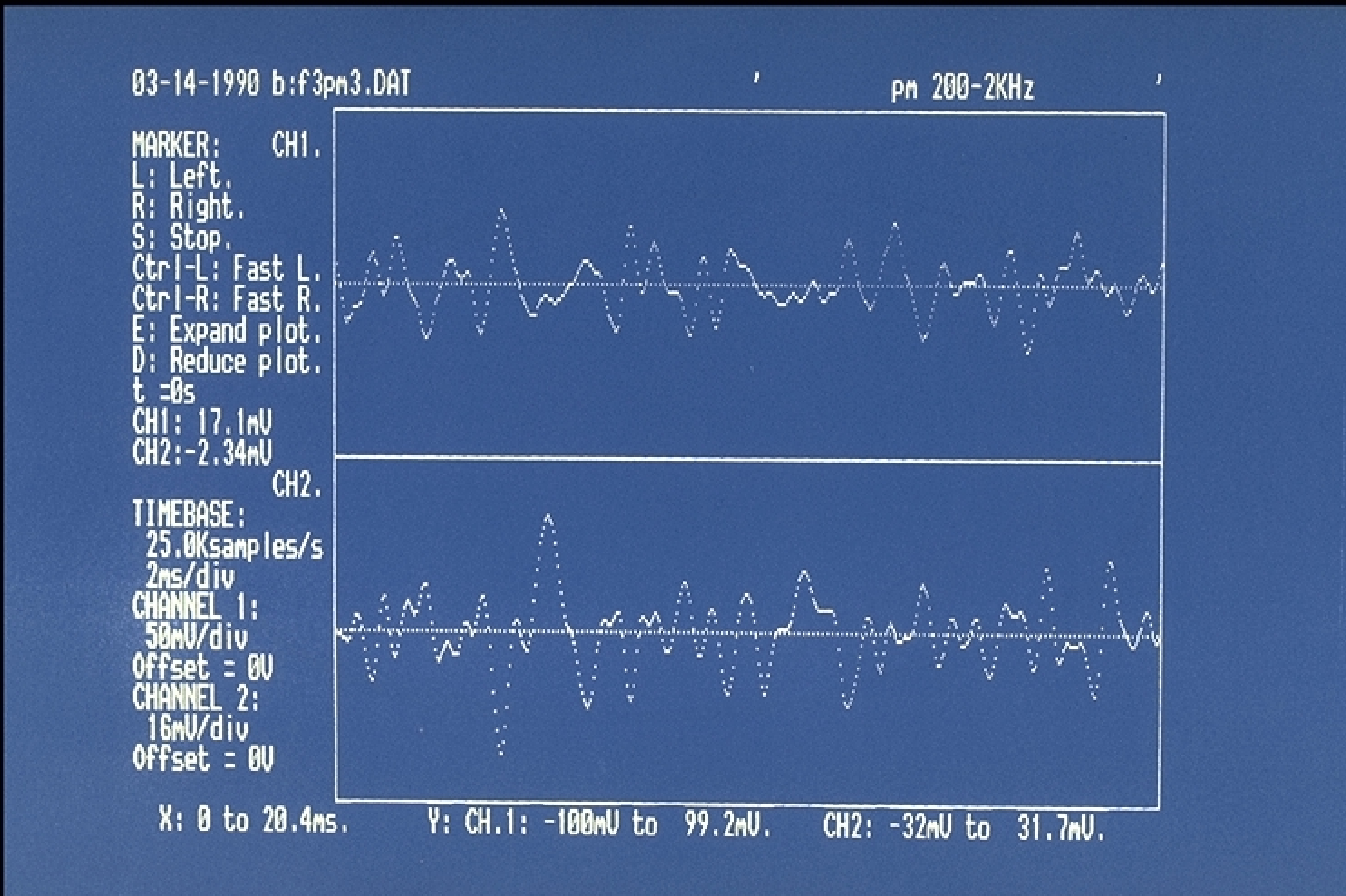
FLICKER NOISE MODEL

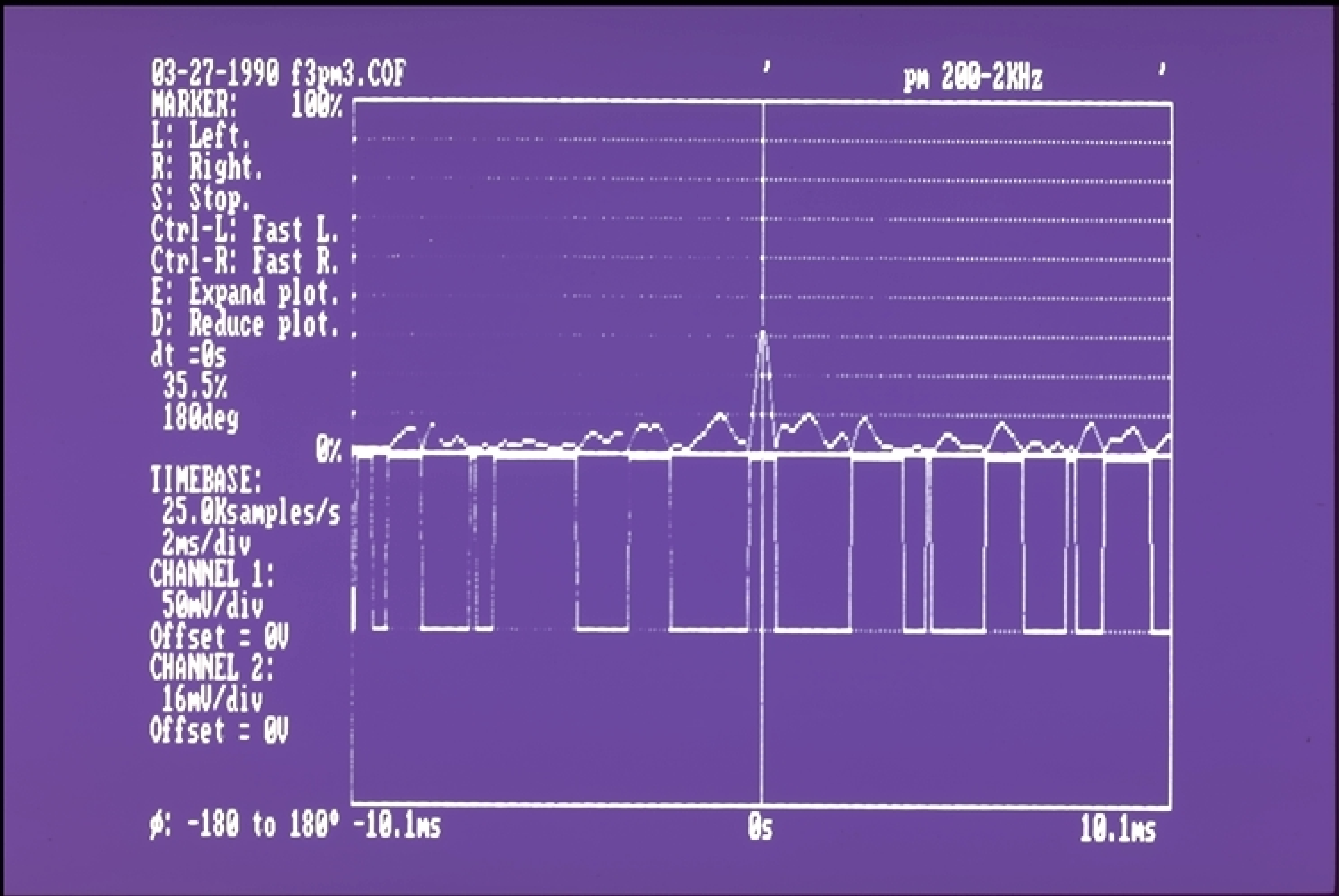
FLICKER NOISE MEASUREMENT SYSTEM



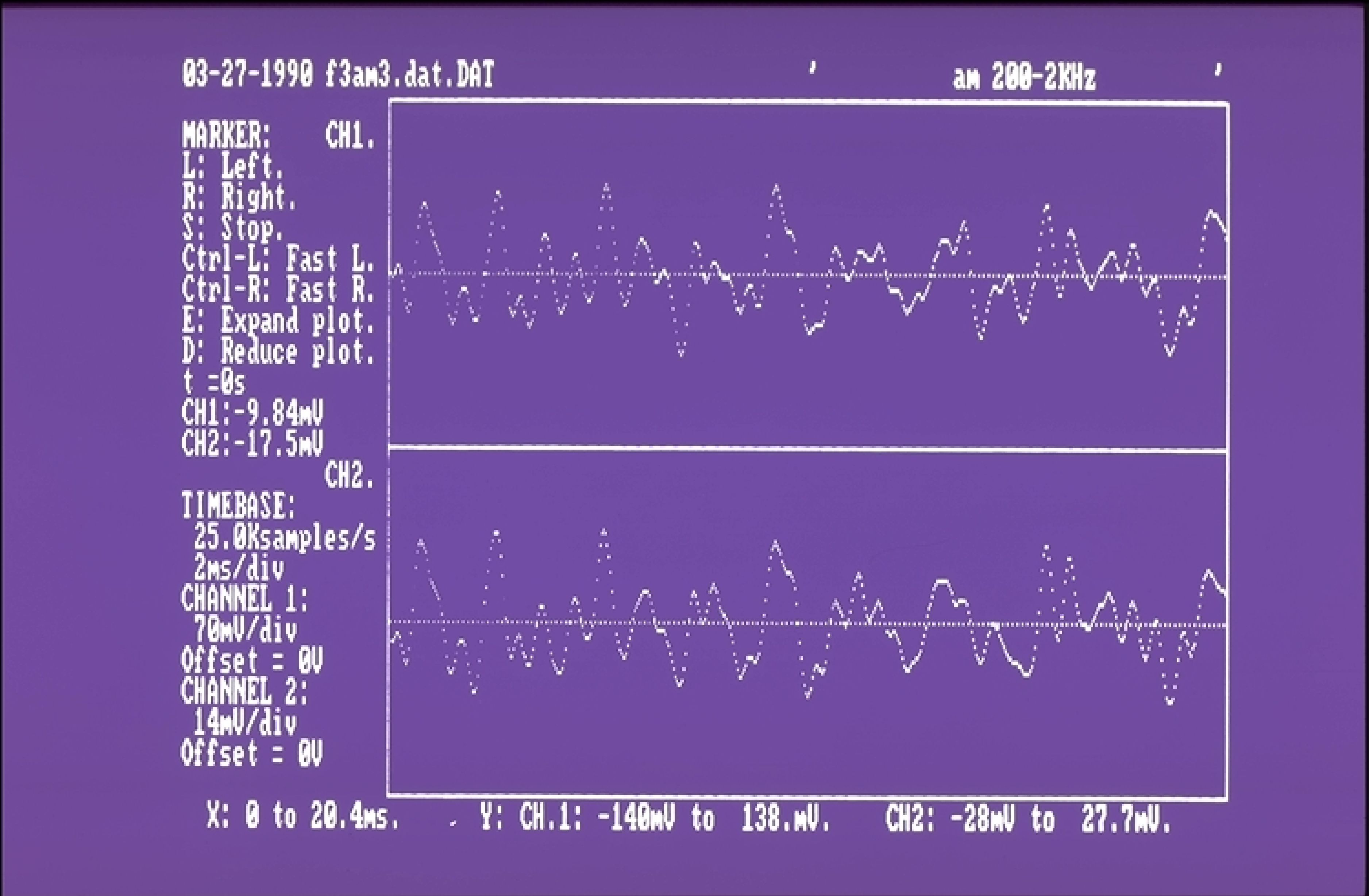

ACME

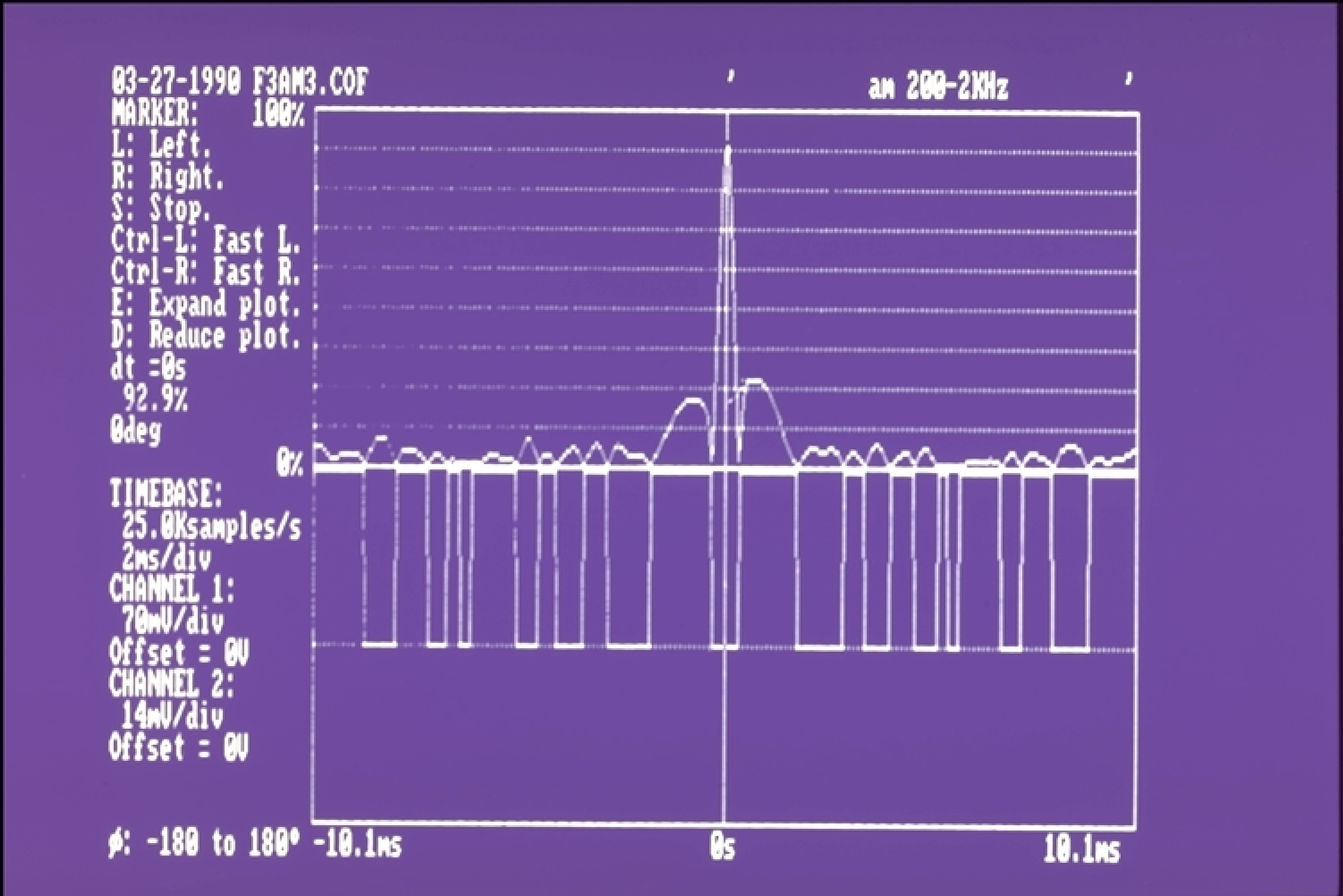

Control
Sweep
Cont.
Single
Start
Stop
Status
Trace
????
sec / div
650.000m
V / div
Trigger
Ch. 1
Ch. 3
Free Run
Rising
Falling
Level
Horizontal Shift
1 dB Range
D 1KHz - 10KHz
C 100Hz - 1KHz
B 10Hz - 100Hz
A 1Hz - 10Hz
Signal Generator
Output 1
Output 2


03-14-1990 b:f3pm3.DAT
pm 200-2KHz
MARKER: CH1.
L: Left.
R: Right.
S: Stop.
Ctrl-L: Fast L.
Ctrl-R: Fast R.
E: Expand plot.
D: Reduce plot.
t =0s
CH1: 17.1mV
CH2:-2.34mV
CH2.
TIMEBASE:
25.0Ksamples/s
2ms/div
CHANNEL 1:
50mV/div
Offset = 0V
CHANNEL 2:
16mV/div
Offset = 0V
X: 0 to 20.4ms.
Y: CH.1: -100mV to 99.2mV.
CH2: -32mV to 31.7mV.



03-27-1990 f3pm3.COF
pm 200-2KHz
MARKER:
100%
L: Left.
R: Right.
S: Stop.
Ctrl-L: Fast L.
Ctrl-R: Fast R.
E: Expand plot.
D: Reduce plot.
dt =0s
35.5%
180deg
0%
TIMEBASE:
25.0Ksamples/s
2ms/div
CHANNEL 1:
50mV/div
Offset = 0V
CHANNEL 2:
16mV/div
Offset = 0V
φ: -180 to 180°
-10.1ms
0s
10.1ms


03-27-1990 f3am3.dat.DAT
am 200-2KHz
MARKER: CH1.
L: Left.
R: Right.
S: Stop.
Ctrl-L: Fast L.
Ctrl-R: Fast R.
E: Expand plot.
D: Reduce plot.
t =0s
CH1:-9.84mV
CH2:-17.5mV
CH2.
TIMEBASE:
25.0Ksamples/s
2ms/div
CHANNEL 1:
70mV/div
Offset = 0V
CHANNEL 2:
14mV/div
Offset = 0V
X: 0 to 20.4ms.
Y: CH.1: -140mV to 138.mV.
CH2: -28mV to 27.7mV.



03-27-1990 F3AM3.COF
am 200-2KHz
MARKER: 100%
L: Left.
R: Right.
S: Stop.
Ctrl-L: Fast L.
Ctrl-R: Fast R.
E: Expand plot.
D: Reduce plot.
dt =0s
92.9%
0deg
0%
TIMEBASE:
25.0Ksamples/s
2ms/div
CHANNEL 1:
70mV/div
Offset = 0V
CHANNEL 2:
14mV/div
Offset = 0V
ϕ: -180 to 180°
-10.1ms
0s
10.1ms

CURRENT METHODS FOR TRANSPOSED FLICKER NOISE REDUCTION

1. Direct LF reduction
2. RF Detection and LF Cancellation
3. Transposed Gain Amplifiers
4. Feedforward Amplifiers

DIRECT LF REDUCTION

Noise reduction was discussed by Riddle and Trew, 1985, who designed the amplifier using a pair of FETS operated in push pull at the microwave frequency but operated in parallel at low frequencies via a low frequency connection between the two bias networks.

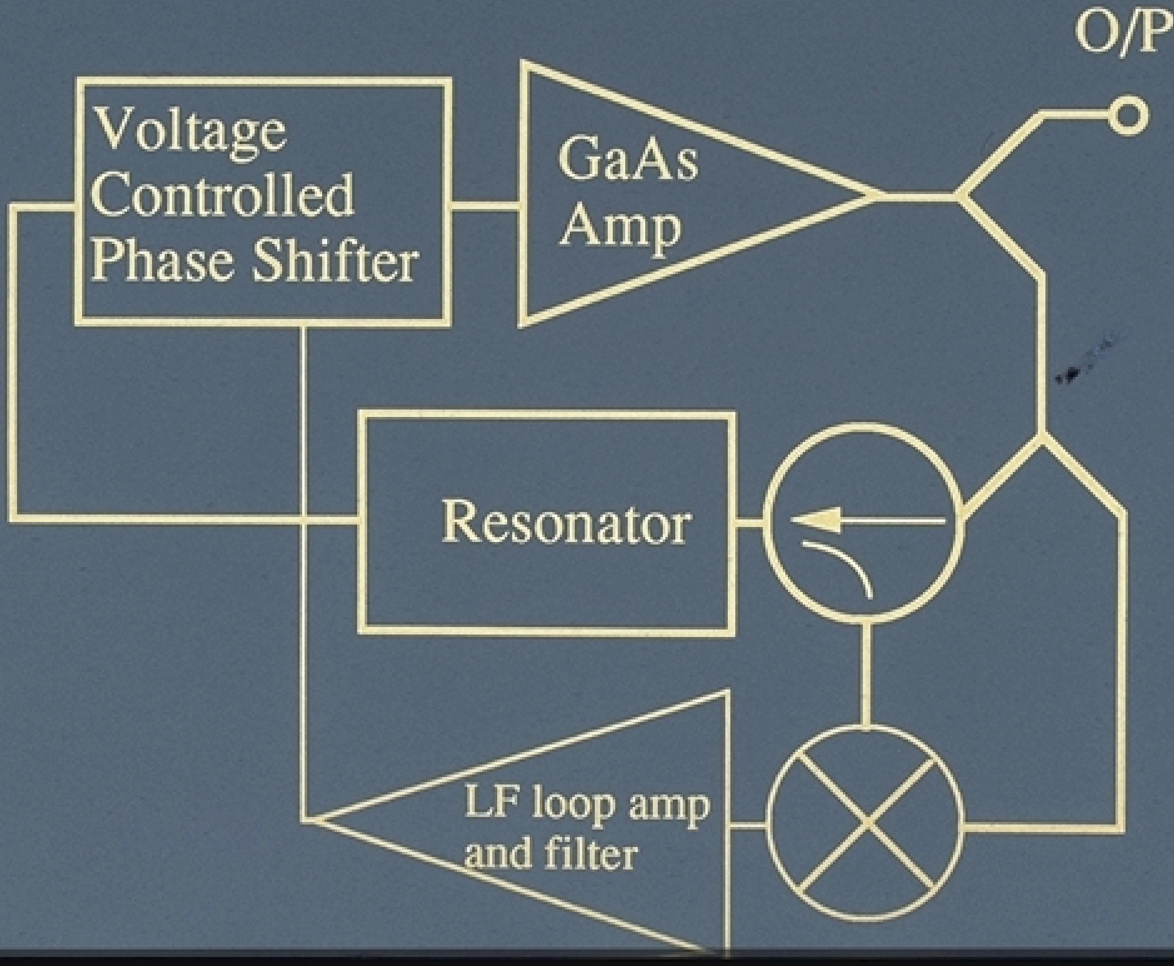
Pringent and Obregon, 1987, used a bias network with a low frequency negative feedback. This reduced the device gain at low frequencies and at the same time reduced the baseband and transposed flicker noise. This assumed that the majority of the Flicker noise was generated by a gate noise source modulating the input non linear capacitor of the GaAs Fet.

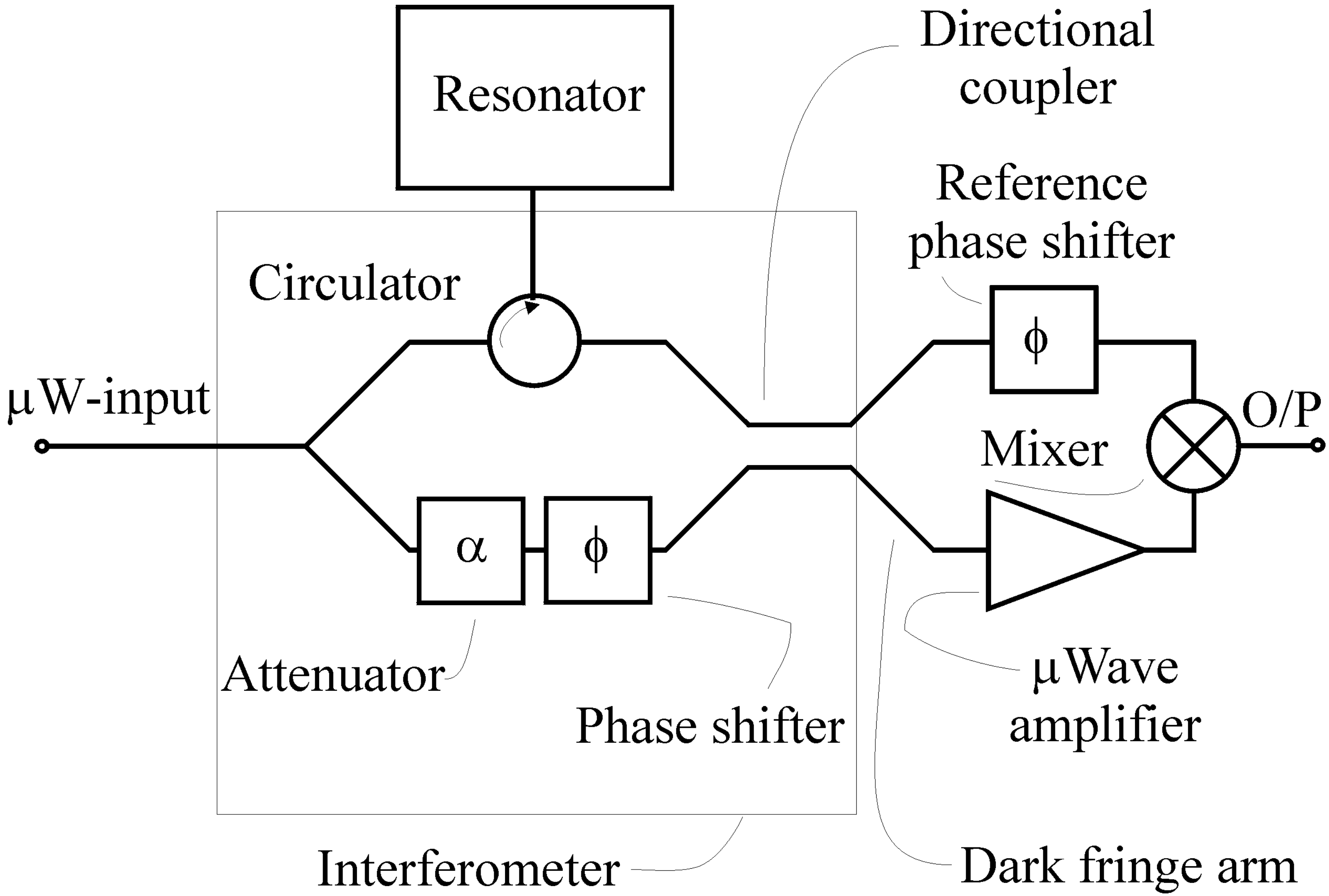
An elegant implementation of the same idea was produced by Mizukami et al 1988 who developed a GaAs mmic in which the impedance presented to the source was arranged to rise at low frequencies. This method would be more difficult to implement with discrete FETs as the parasitics need to be very low indeed.

These methods have all reduced the low frequency flicker noise present at the device terminals, but this often does not necessarily correlate well with the oscillator flicker noise reduction. The transposed flicker noise depends on the nature of the internal noise sources, and the transposition mechanism. All of these vary greatly between device manufacturers.

Flicker Noise Reduction in GaAs Oscillators

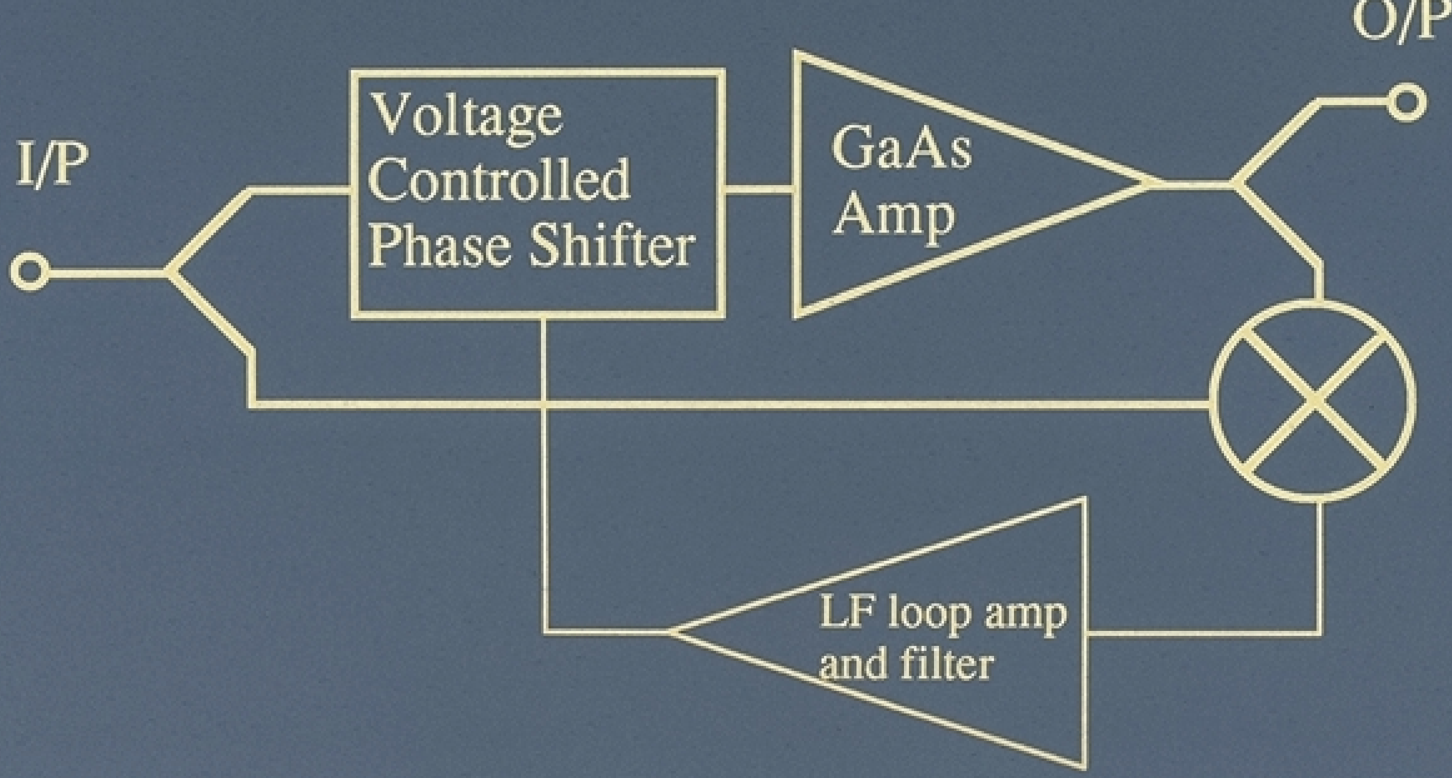
Z.Galani , M.J. Bianchini, R.C. Waterman, R. Dibiase,
R.W Laton and J.B. Cole, IEEE Trans. MTT 32 1984



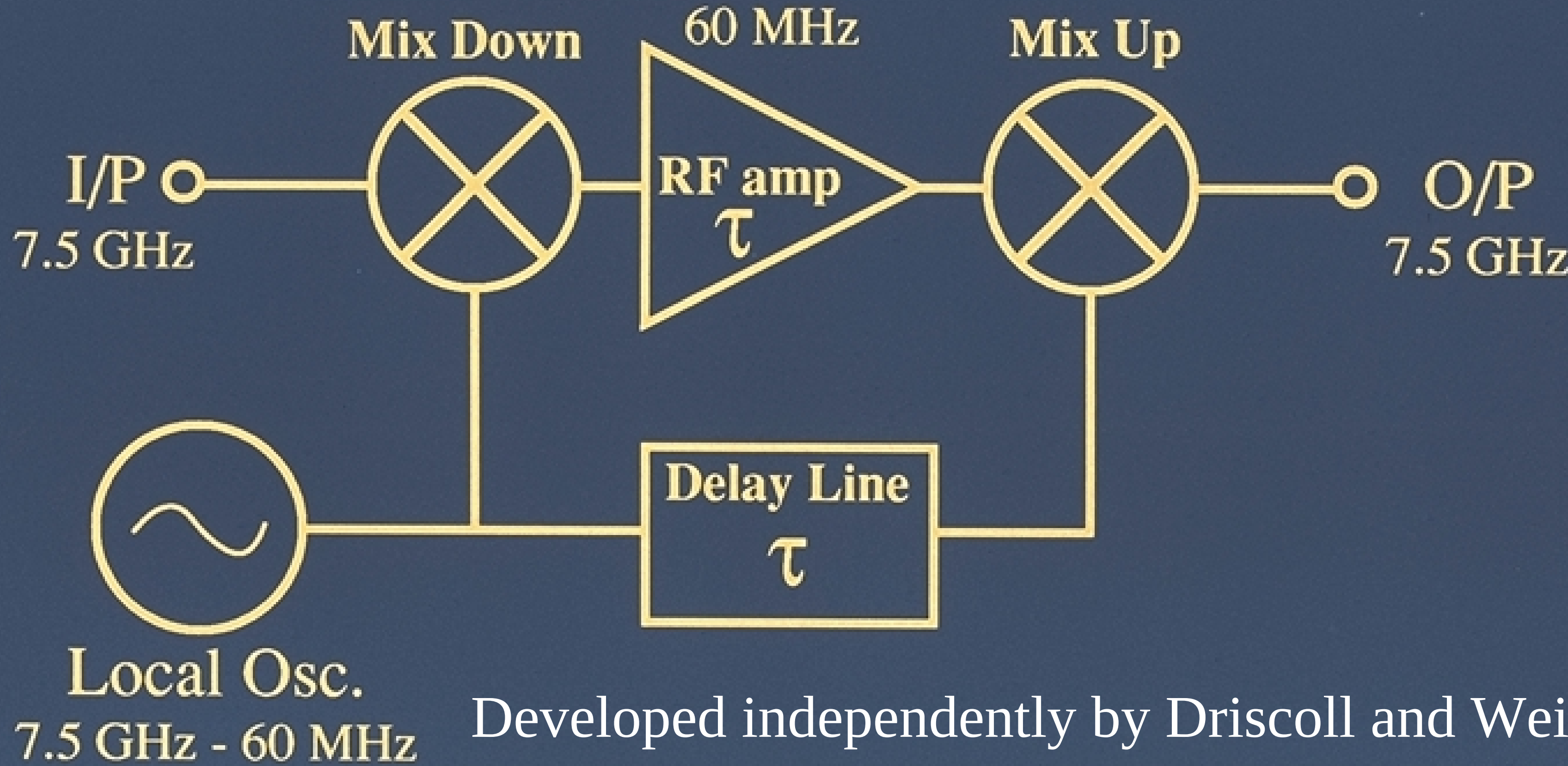


Ultra low noise frequency discriminator for lowest noise X band oscillators - Ivanov, Tobar and Woode [32], [33]

Wide Bandwidth Flicker Noise Reduction in GaAs Amplifiers
M. Driscoll, FCS 1995

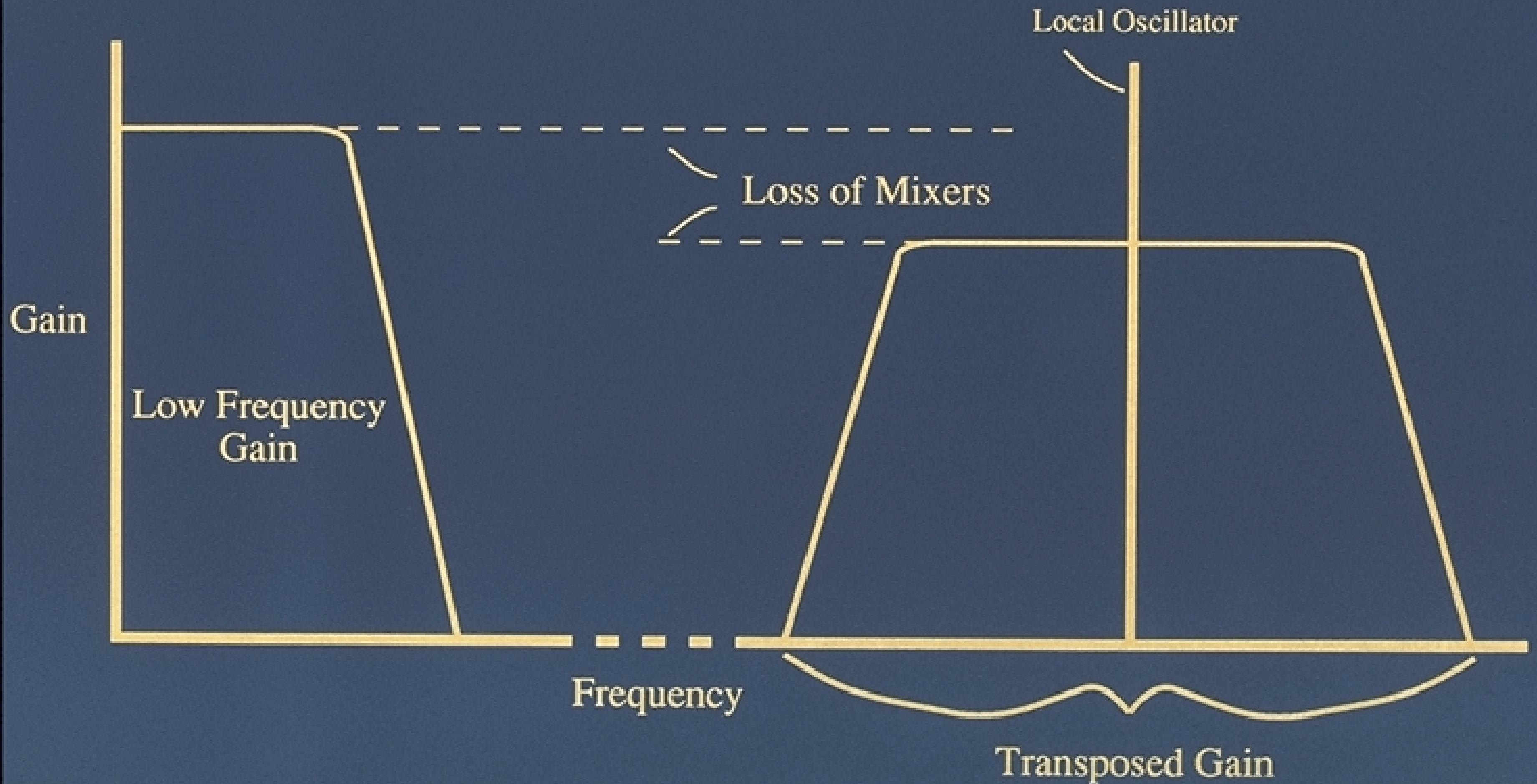


TRANSPOSED GAIN AMPLIFIER

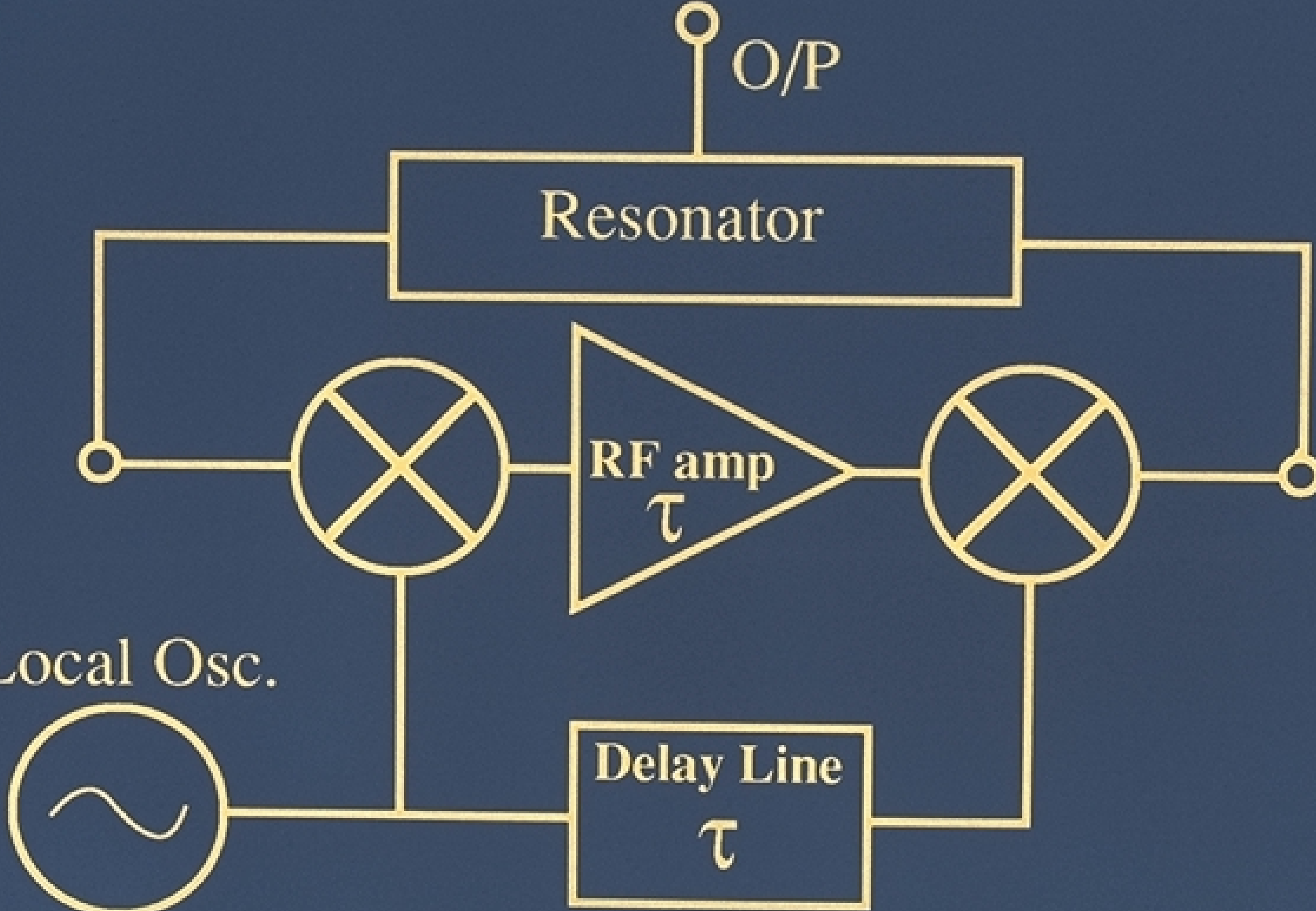


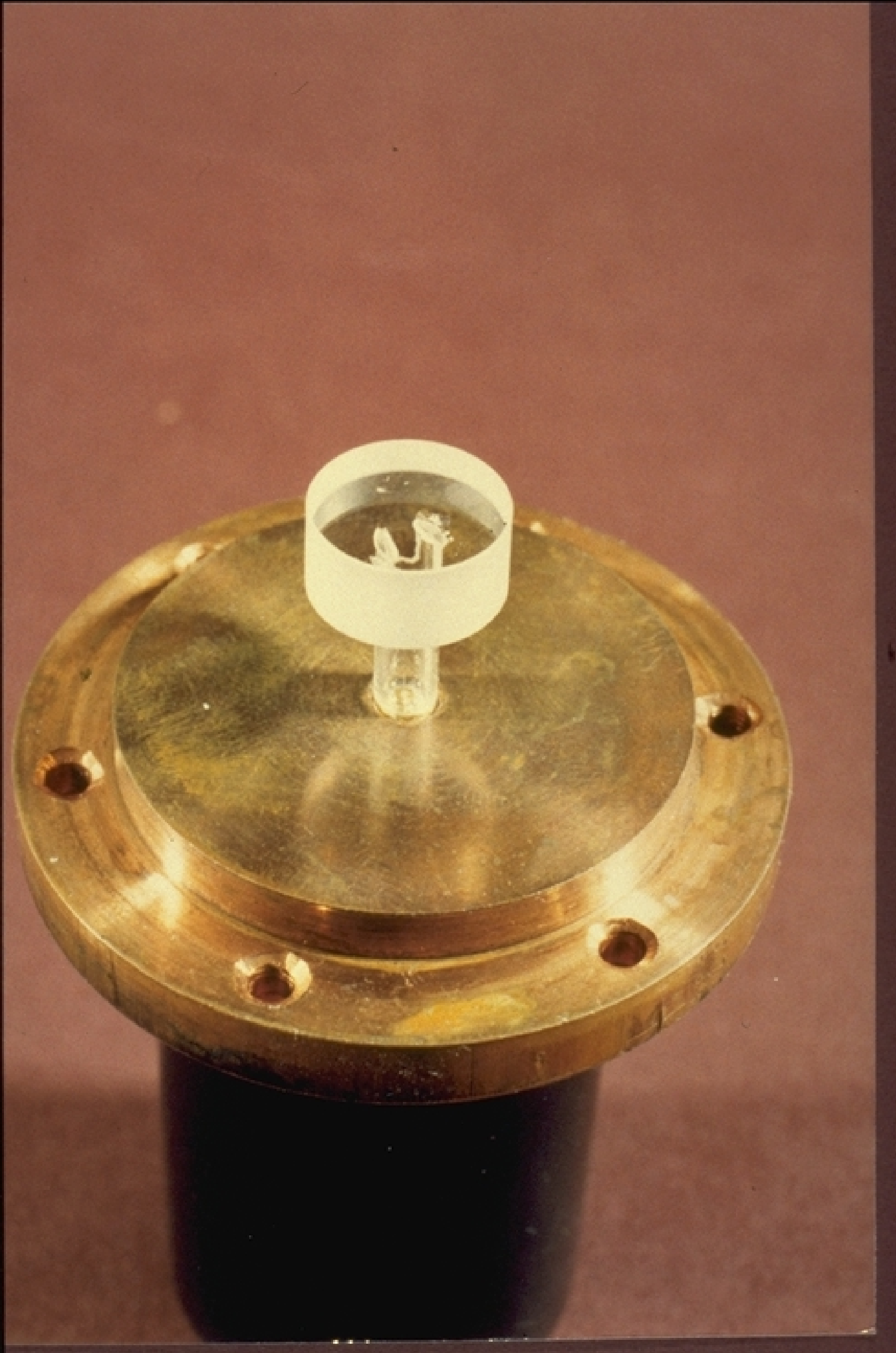
Developed independently by Driscoll and Weinert [11] and Everard and Page-Jones [17-19]

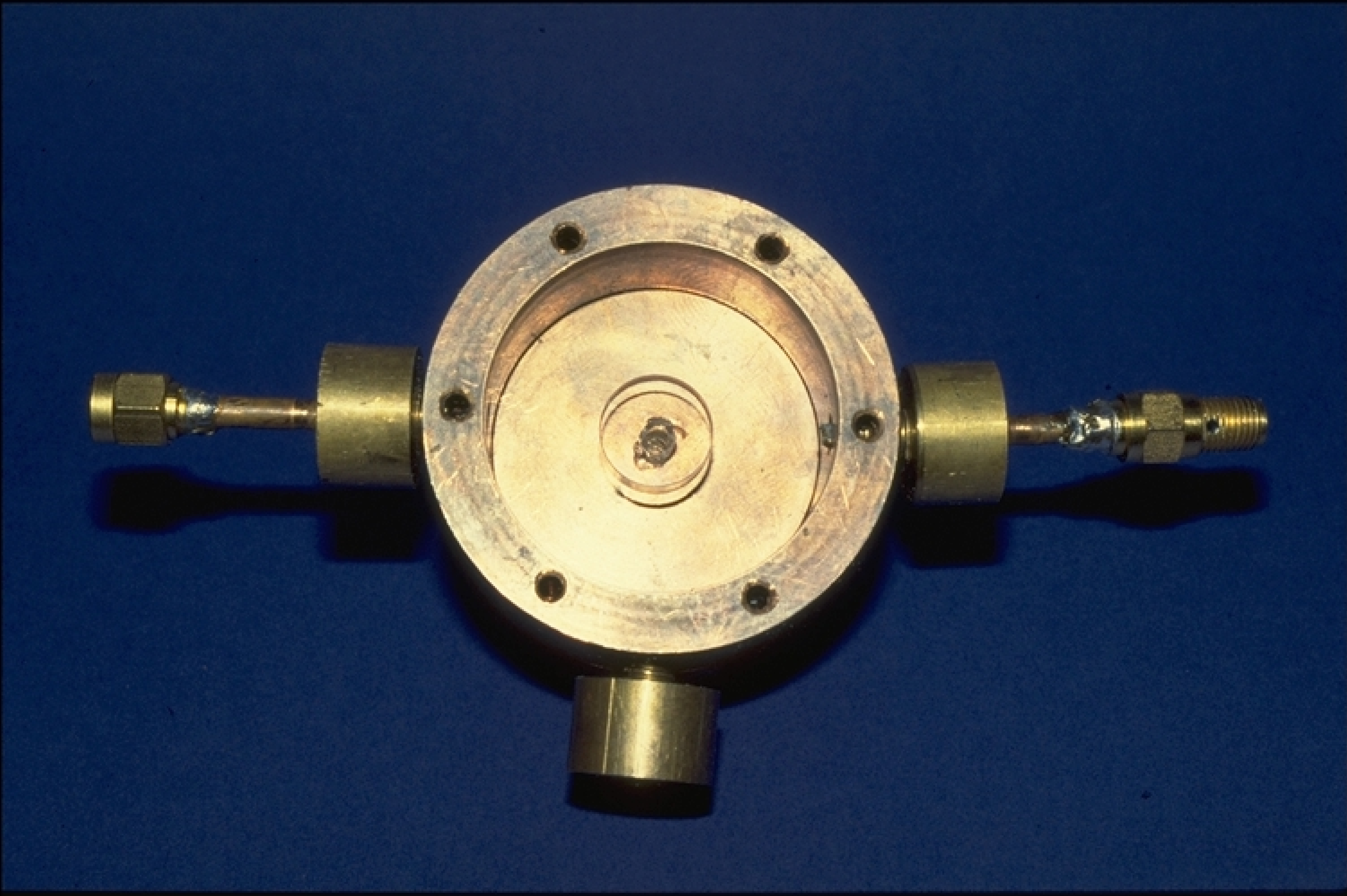
GAIN PRODUCED IN TRANSPOSED GAIN AMPLIFIER

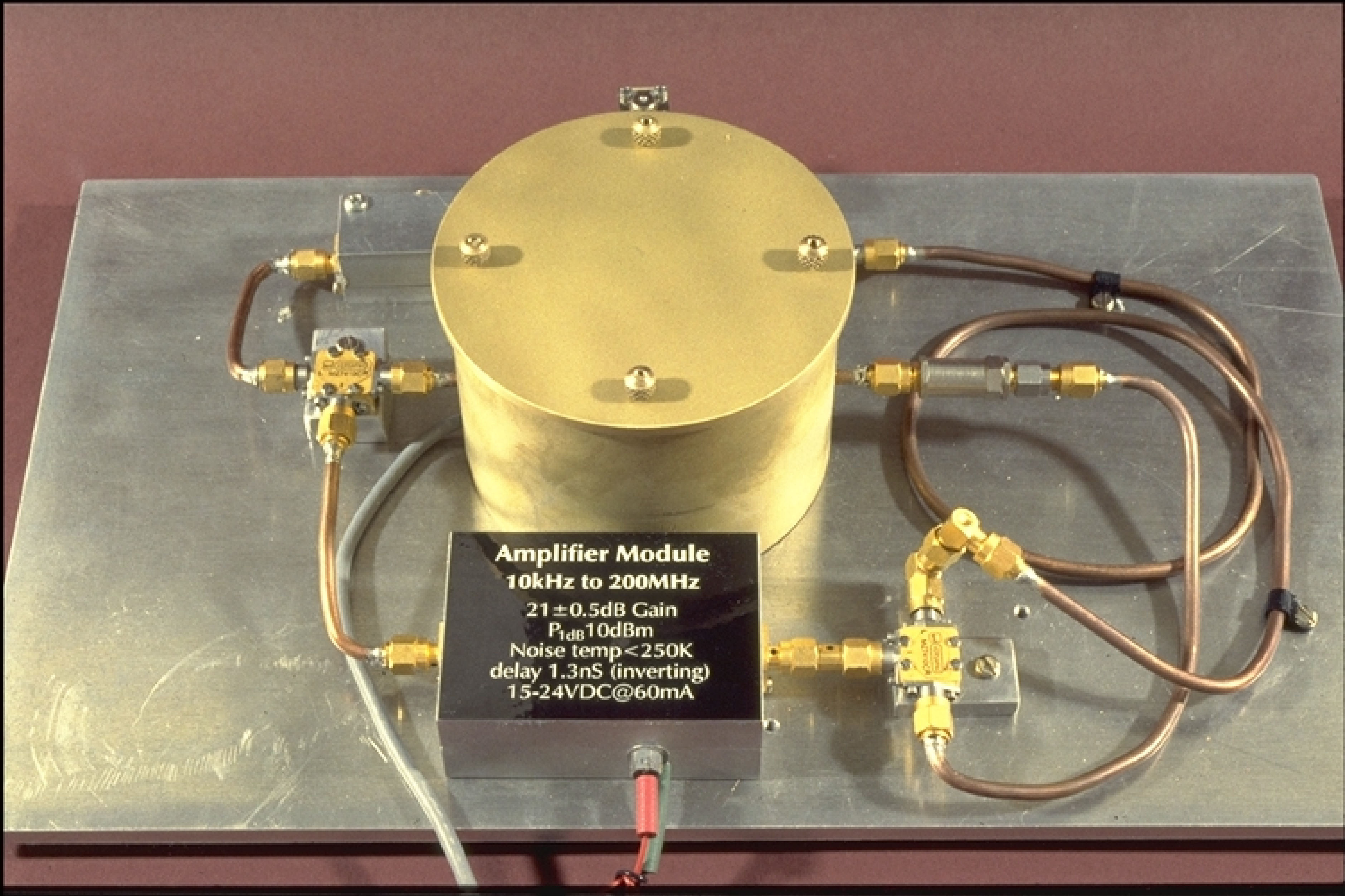


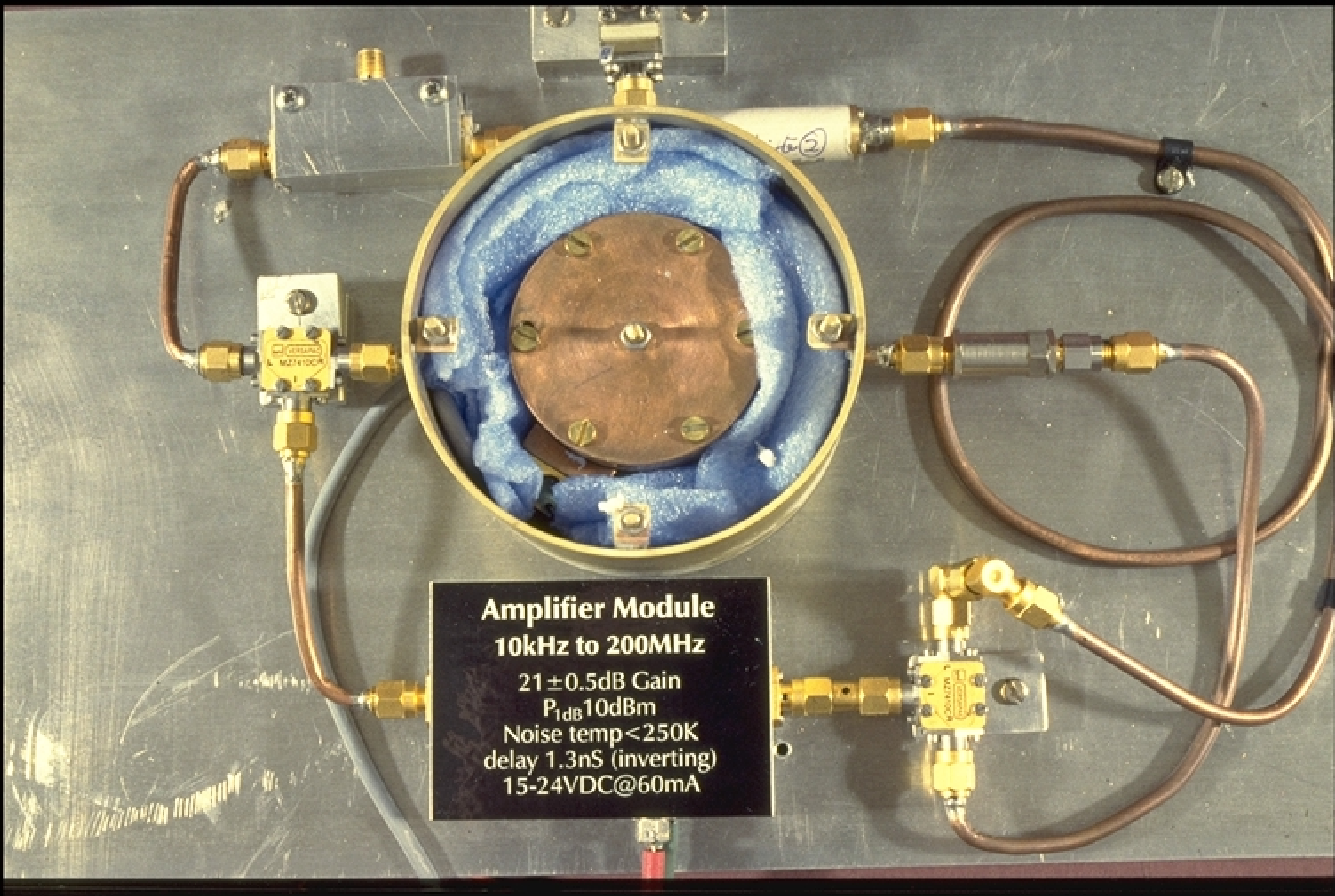
TRANSPOSED GAIN OSCILLATOR








Amplifier Module
10kHz to 200MHz
21±0.5dB Gain
P1dB10dBm
Noise temp<250K
delay 1.3nS (inverting)
15-24VDC@60mA


Amplifier Module
10kHz to 200MHz
21±0.5dB Gain
P1dB10dBm
Noise temp<250K
delay 1.3nS (inverting)
15-24VDC@60mA

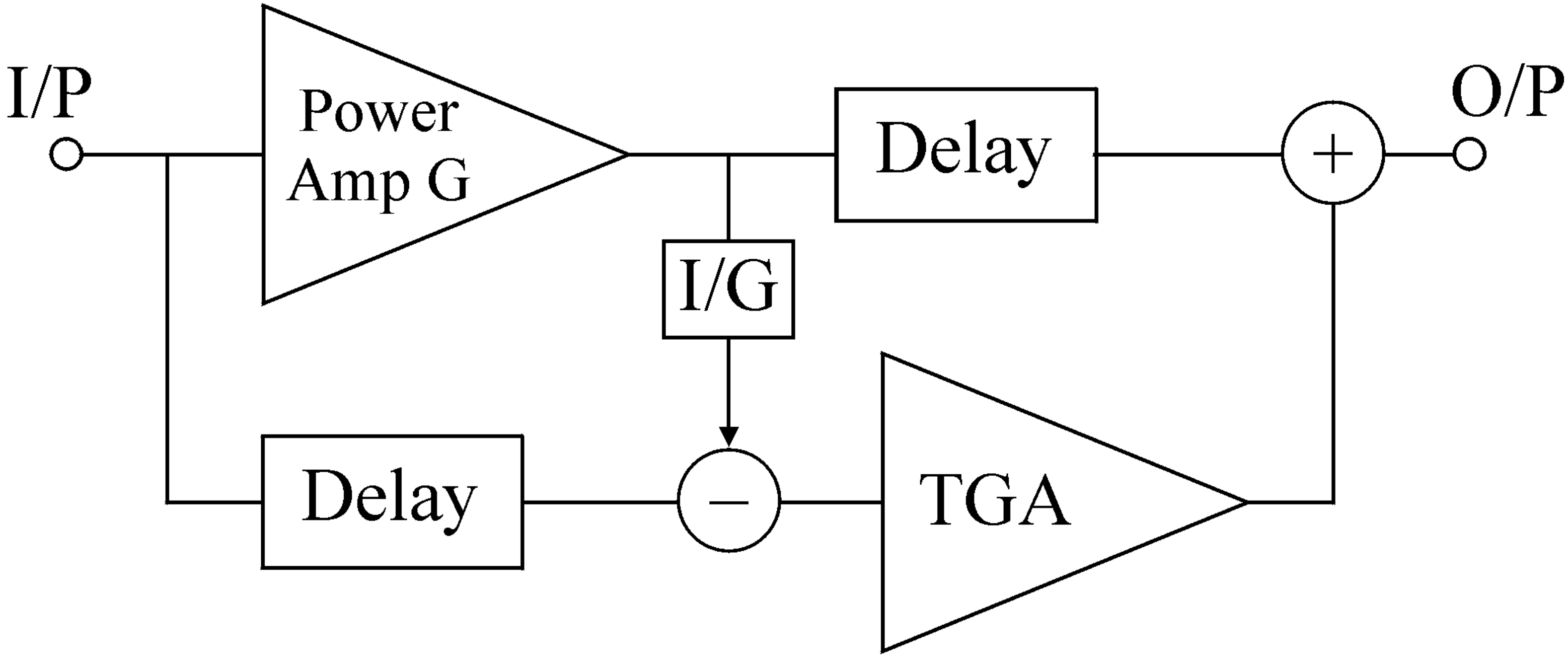
Phase Noise Performance

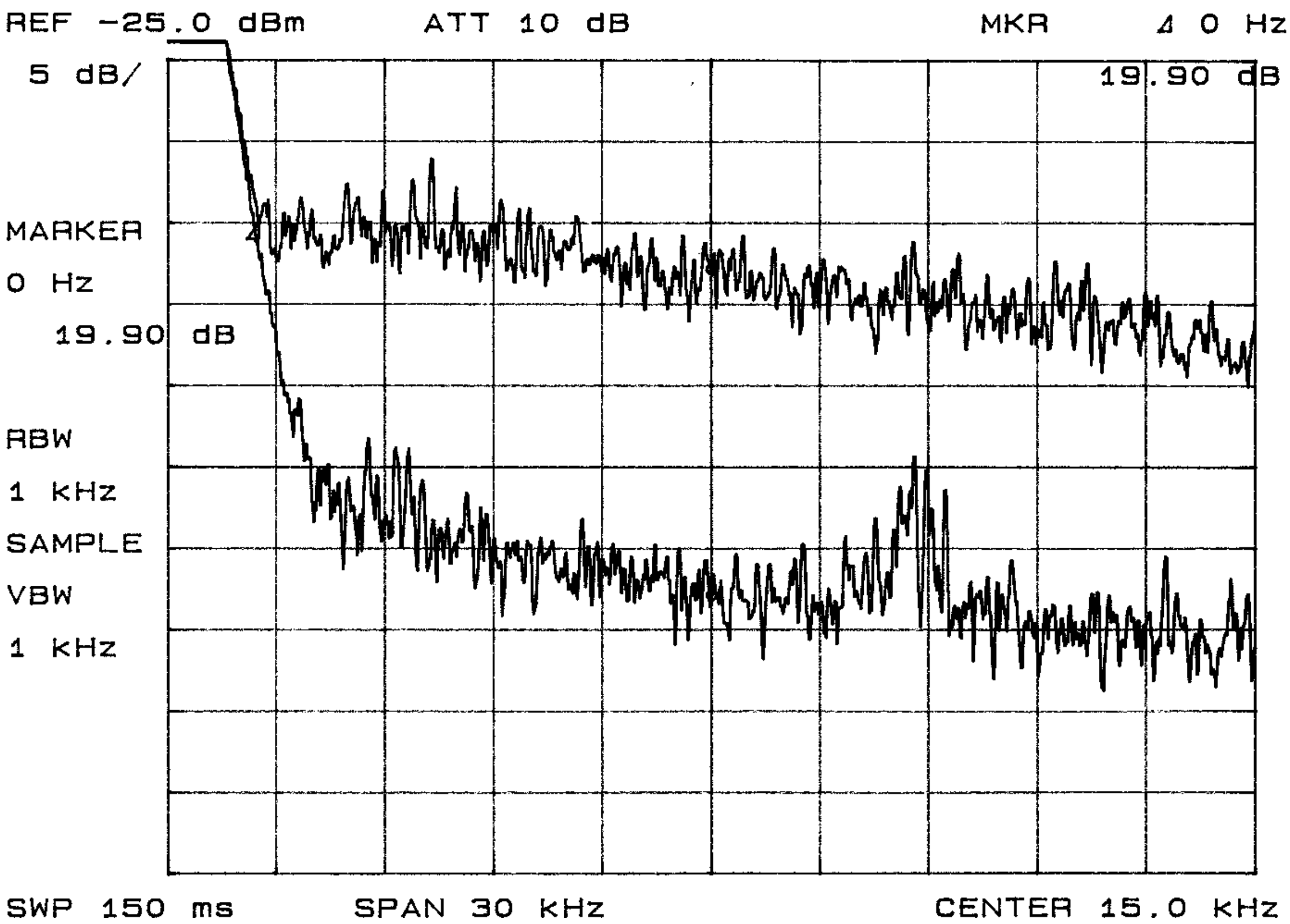
- F_0= 7.6GHz
- Q_0 = 44,000
- P_{AVO} = 8dBm (6.3mW)
- Noise Figure = 15dB including image noise
- Flicker noise corner ~ 1kHz
- L_{FM} = -136dBc@10kHz (theory -139dBc)

Problems with TGO

O/P power max ~ 8dBm NF ~ 15dB

Therefore use FEEDFORWARD

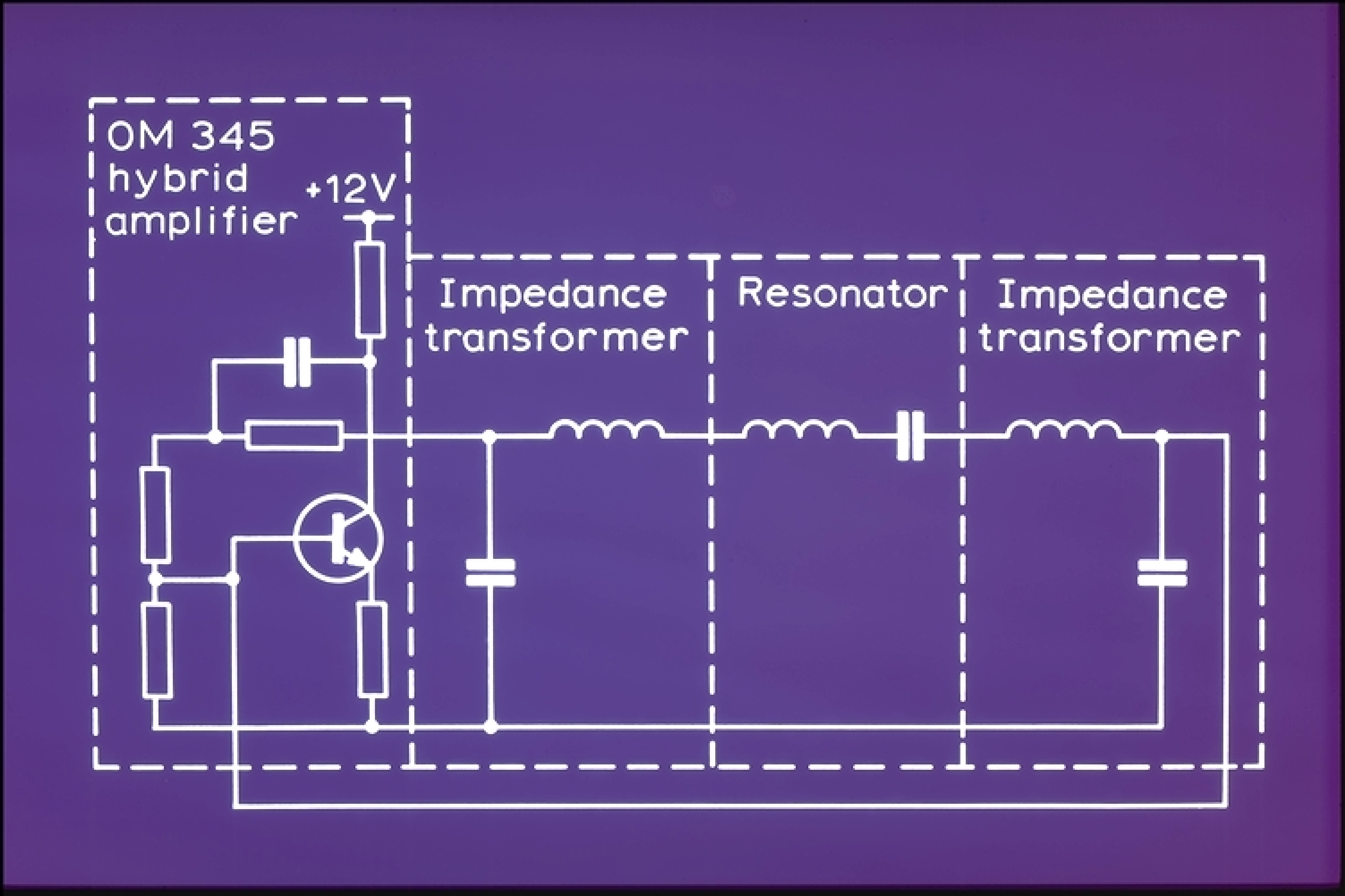




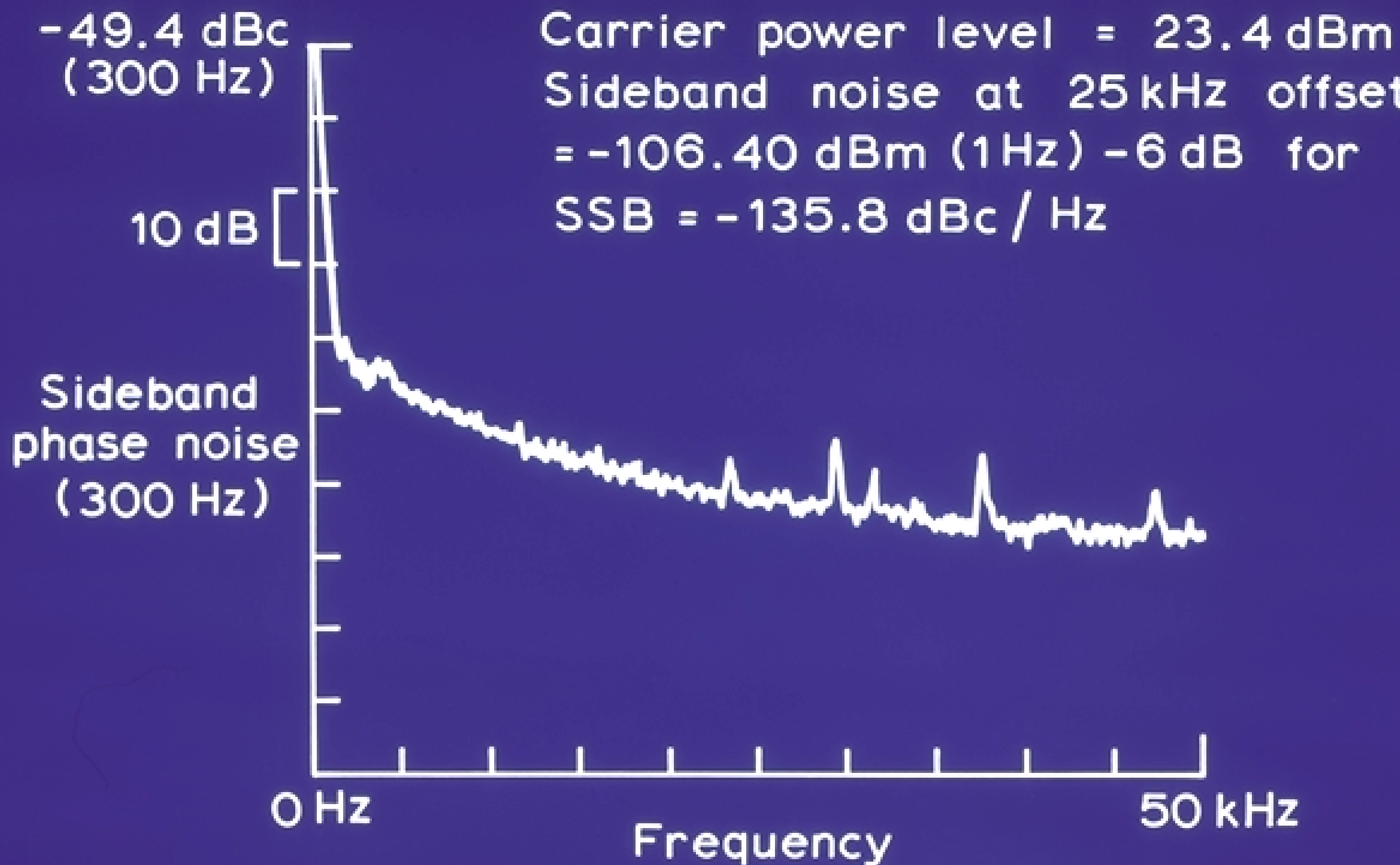
Residual flicker noise reduction in 1 Watt GaAs Feedforward Amp
Broomfield and Everard FCS 2000

Oscillator designs

- LC
- SAW
- Transmission Line
- Helical
- Tuning
- Detailed designs
  - LC
  - Transmission line



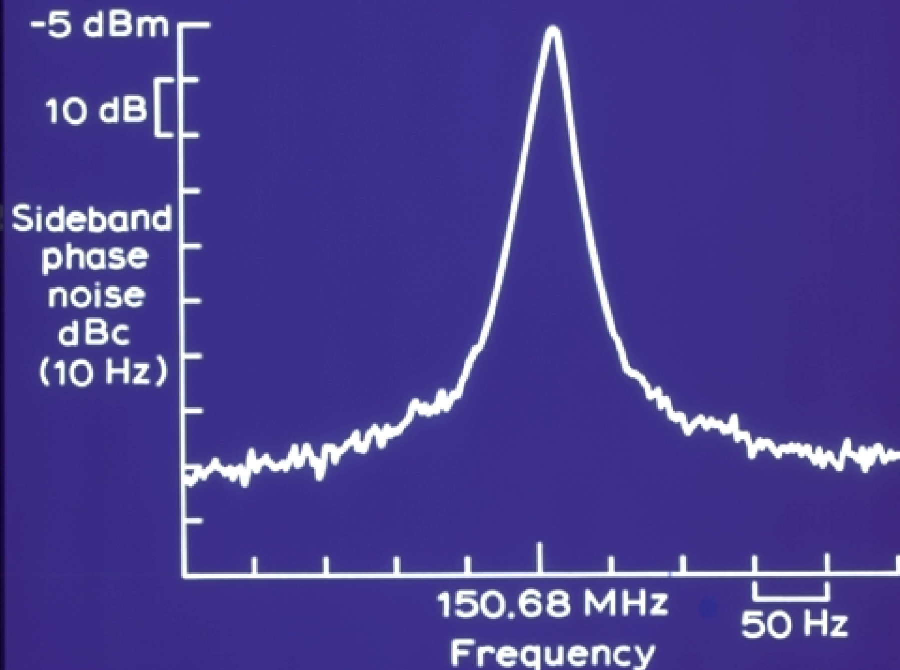
OM 345
hybrid
amplifier
+12V
Impedance
transformer
Resonator
Impedance
transformer



VID AVG 30, RES BW 300 Hz, VBW 300 Hz, SWP 1.0 s.

Oscillator phase noise against frequency.

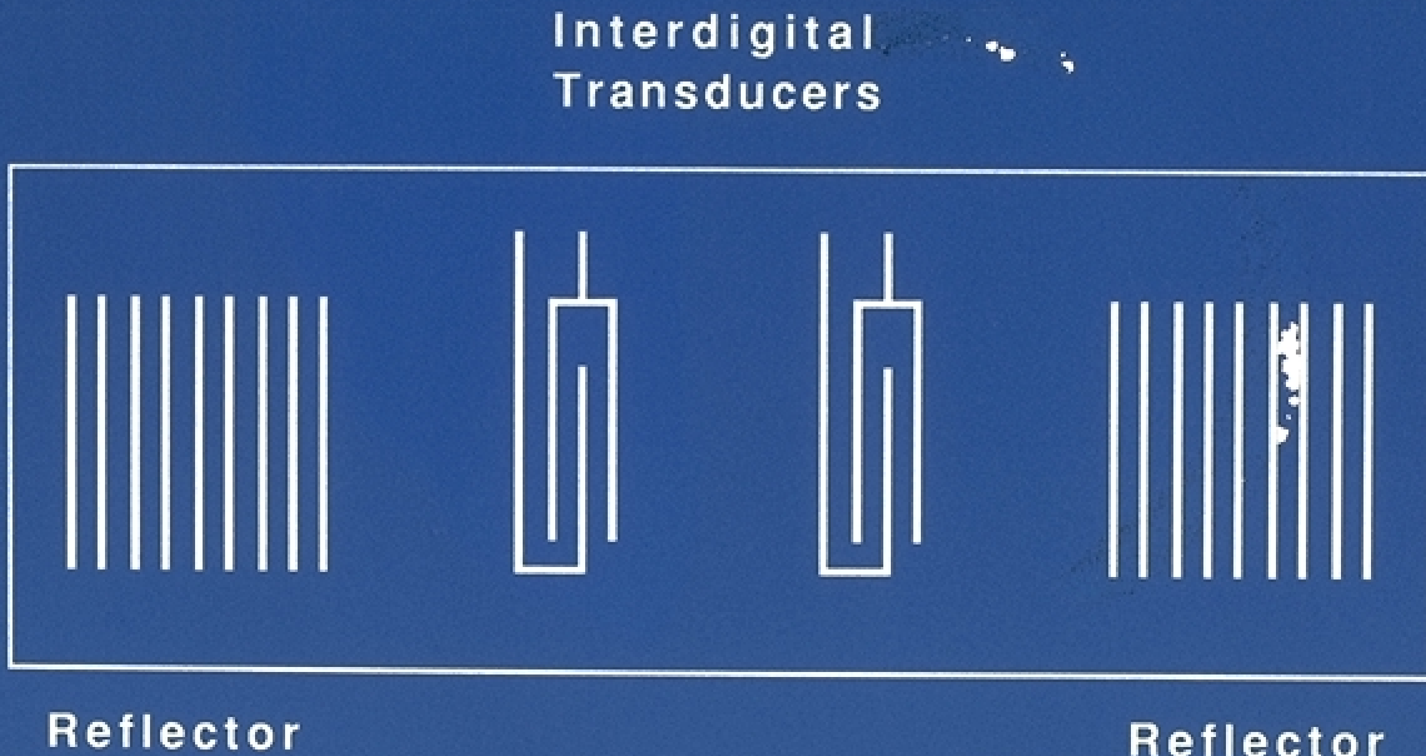
Carrier power level = -5 dBm
Sideband noise at 250 Hz offset
= -92.50 dBm (1 Hz) = -87.5 dBc/Hz



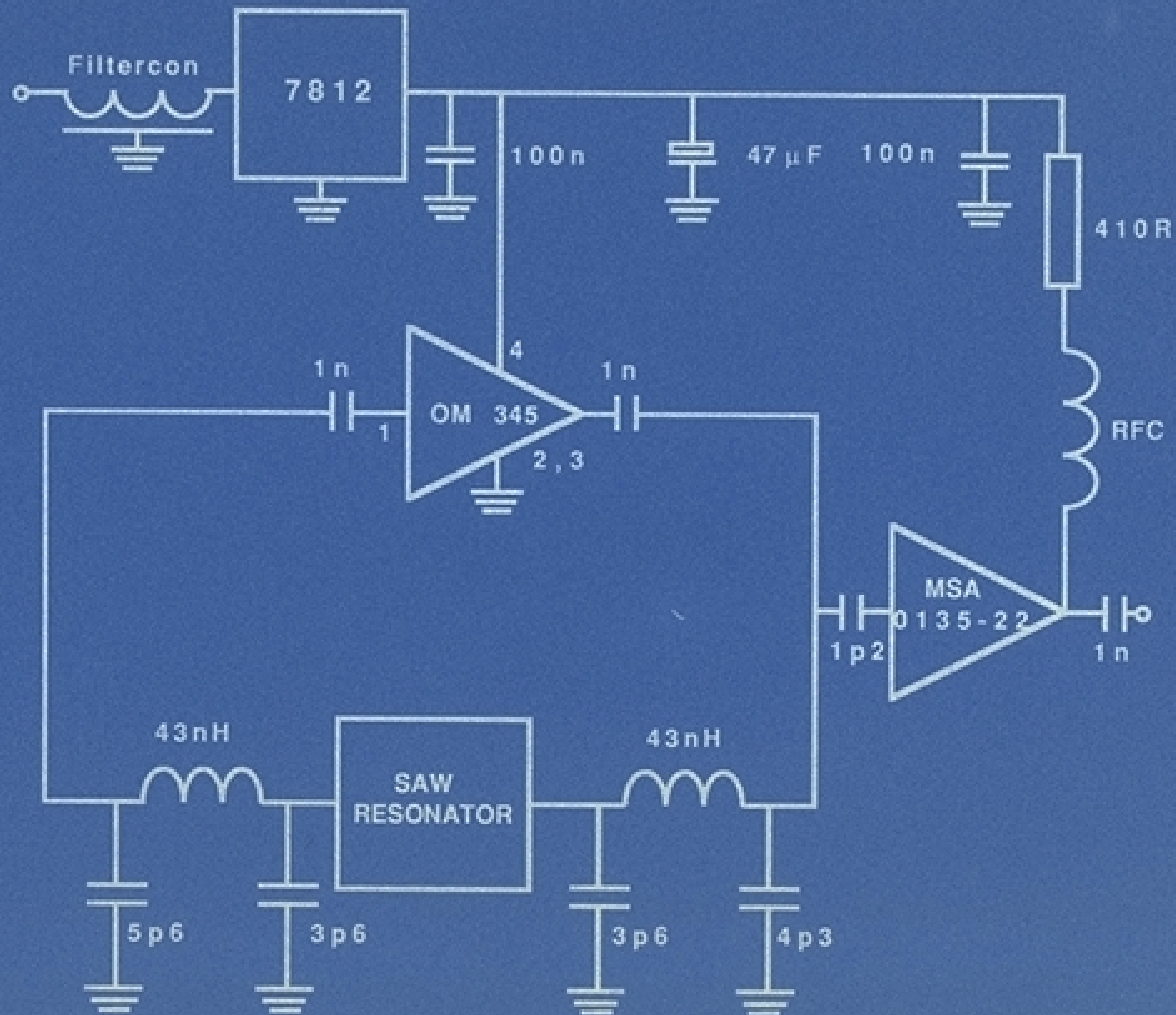
VID AVG 10, RES BW 10 Hz,
VBW 10 Hz, SWP 10s.

Oscillator phase noise against frequency.

SURFACE ACOUSTIC WAVE RESONATOR



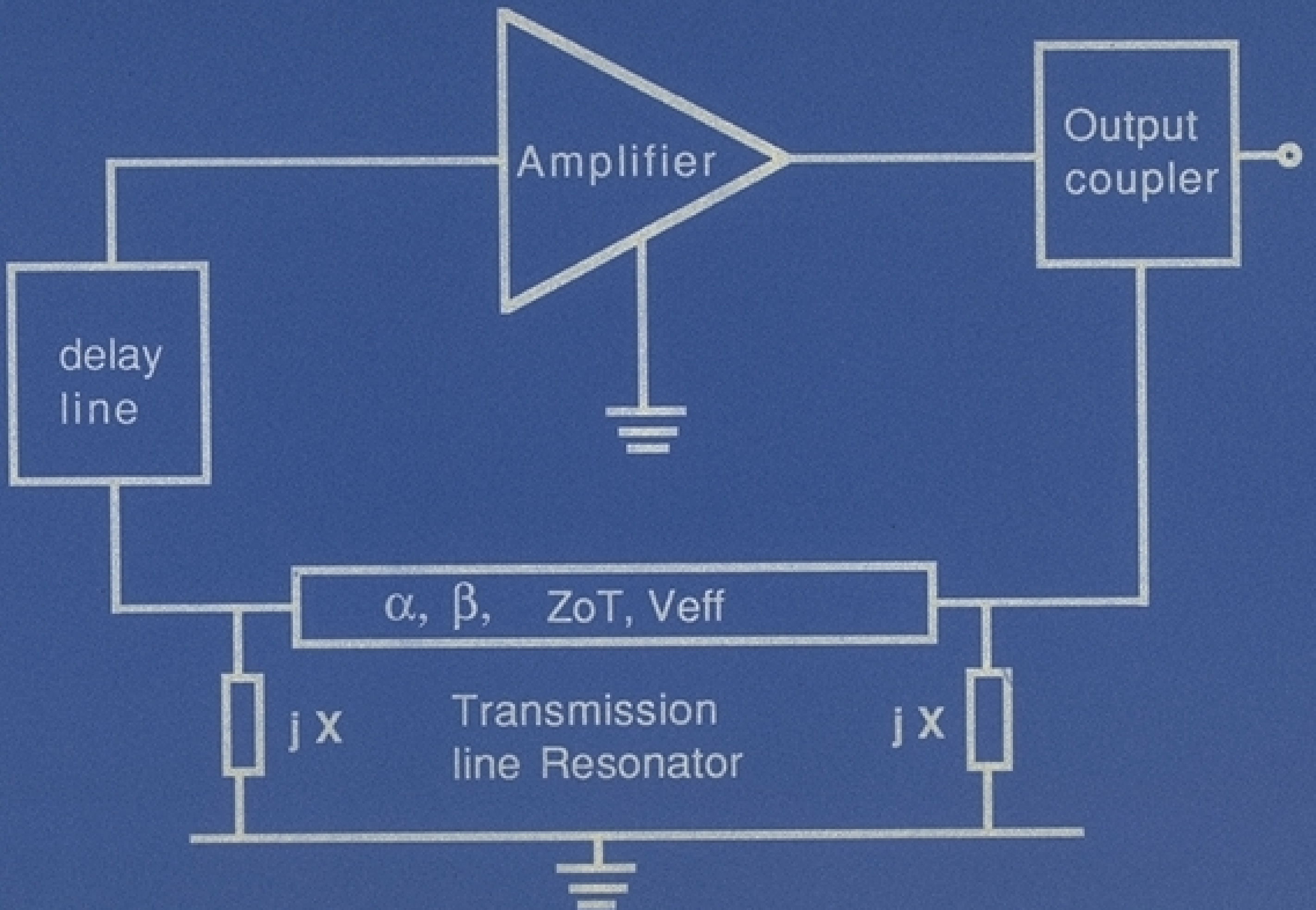
262MHz Surface Acoustic Wave Oscillator



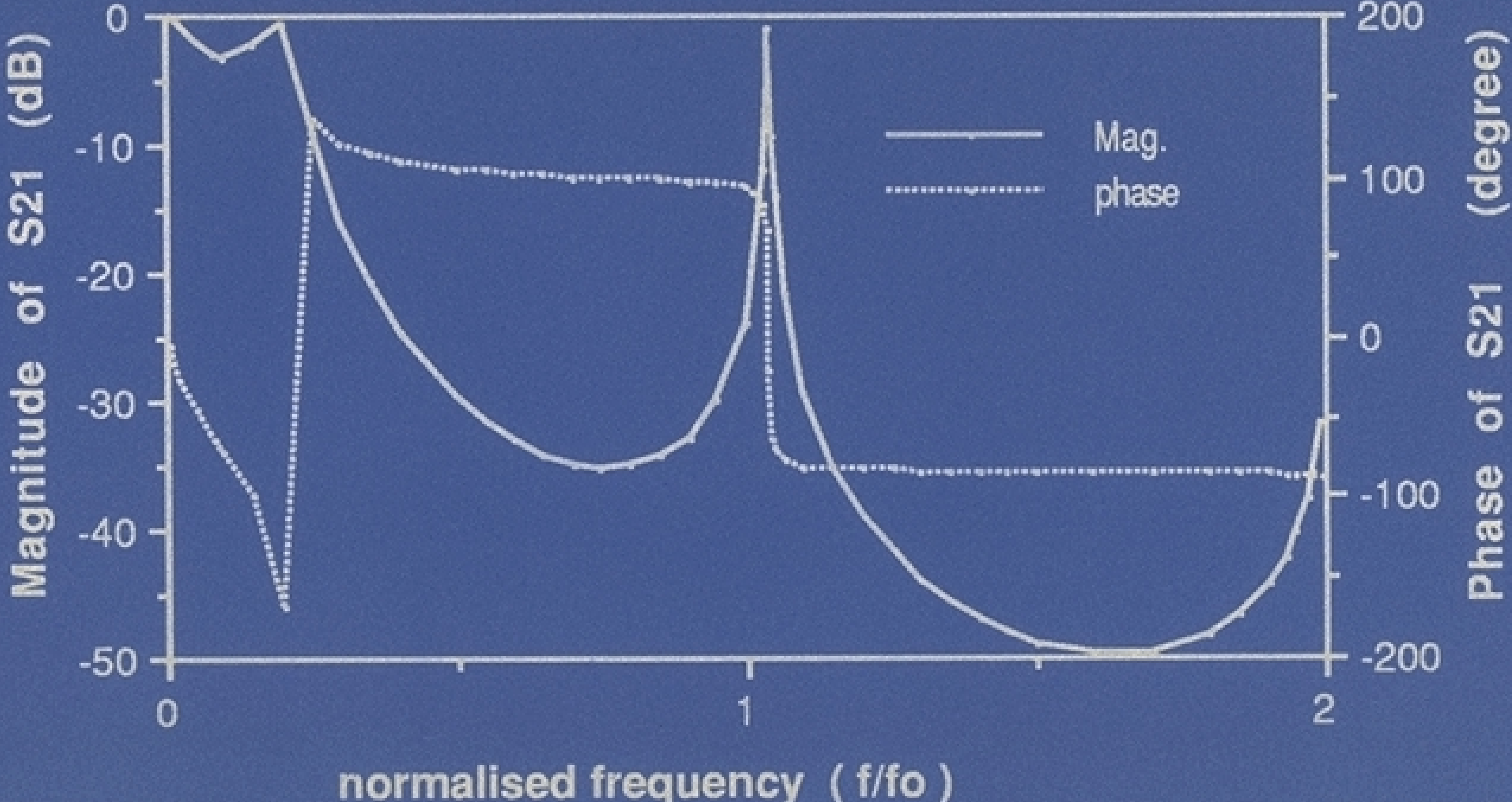
262 MHz SAW Oscillator

- Phase noise performance of -130dBc/Hz at 1kHz offset - limited by measurement
- Montress, Parker, Loboda and Greer [20] demonstrated high power 500MHz SAW oscillators with -140dBc/Hz at 1kHz offset

TRANSMISSION LINE OSCILLATOR



Frequency Response of Resonator



Close to the resonant peak and for small αL (< 0.05) and $\delta f/f_o << 1$,

$$f_o = (v_{eff}/2L)\{ 1 + (1/\pi)\tan^{-1}(2/X)\},$$

f_o = resonant frequency and $\delta f = f - f_o$

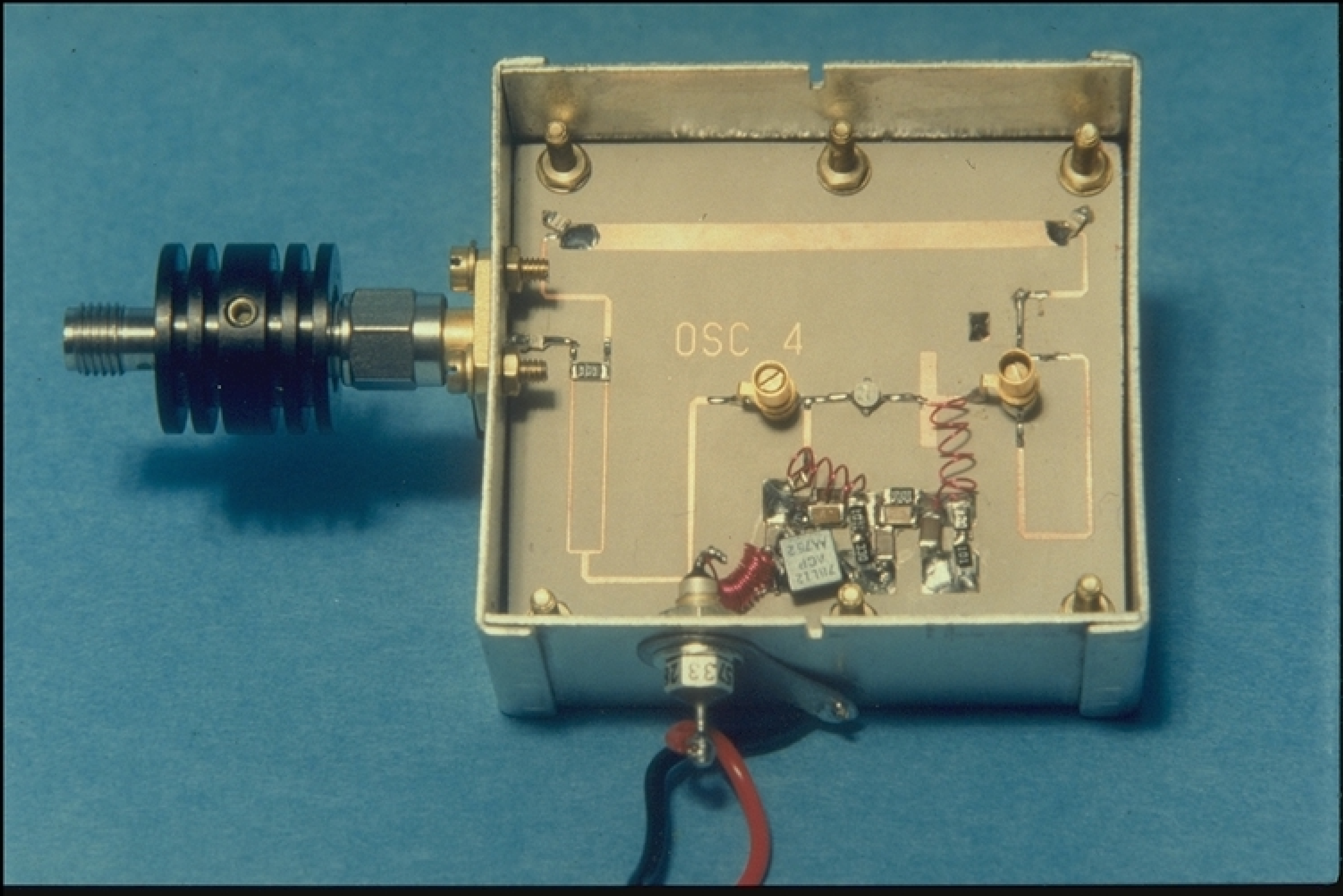
$$S_{21}(\delta f) = S_{21}(0)/\{1 + j2Q_L(\delta f/f_o)\}$$

$$S_{21}(0) = 1/\{1 + (\alpha L/2)X^2\}$$

$$Q_L = \pi S_{21}(0)X^2/4$$

From the last two equations it can be seen that the insertion loss and the loaded Q factor of the resonator are interrelated.

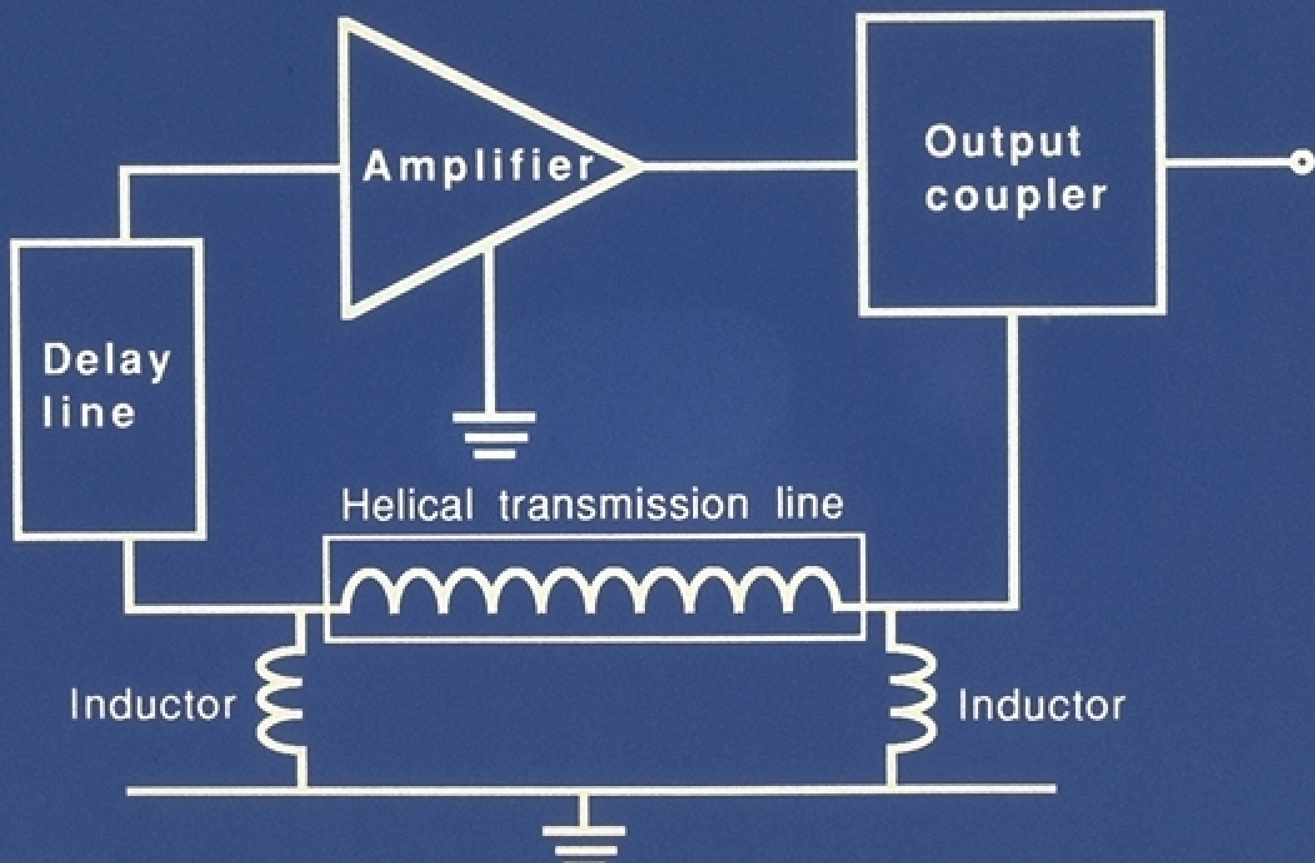
As the shunt capacitors (assumed to be lossless) are increased the insertion loss increases towards infinity and Q_L increases to a limiting value of $\pi/2\alpha L$ which we will define as Q_o.


OSC 4

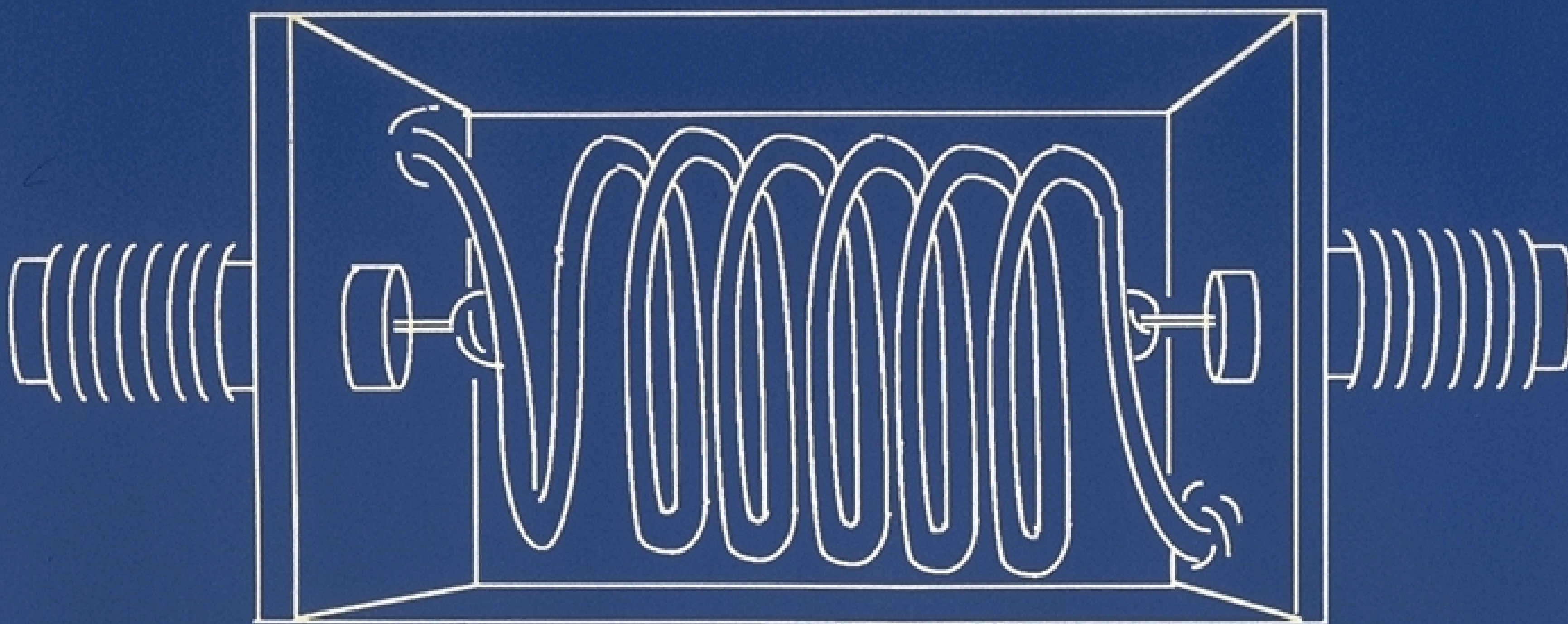
Noise performance of 1.5GHz Osc.

- Q_0 = 83, αl = 0.019 , substrate ε_r10
- O/P power = 3.1dBm
- Noise Figure = 3dB
- Noise performance = -104dBc/Hz @ 10kHz
- Within 2dB of the theory

HELICAL RESONATOR OSCILLATOR



HELICAL RESONATOR



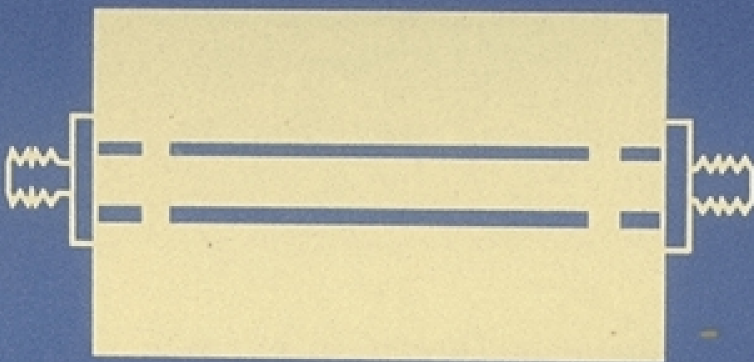
Helical Resonator

Helical resonator oscillators [22]

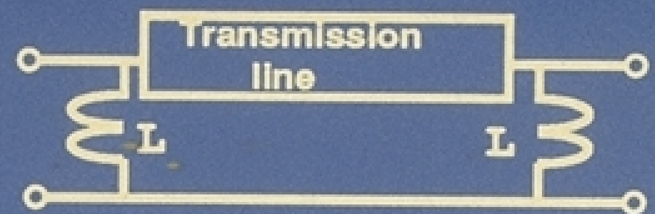
- 900 MHz, Q_0 = 582
- O/P power = 0dBm
- Noise Factor = 6dB
- Noise performance =
- -127dBc/Hz @ 25kHz
- Within 2dB of theory

- 1.6GHz, Q_0 = 382
- O/P power = 0dBm
- Noise Figure = 3dB
- Noise performance =
- -120dBc/Hz @ 25kHz
- Within 2dB of theory

Z_0 of helix = 340Ω measured using
Time Domain Reflectometry



STRIPLINE RESONATOR

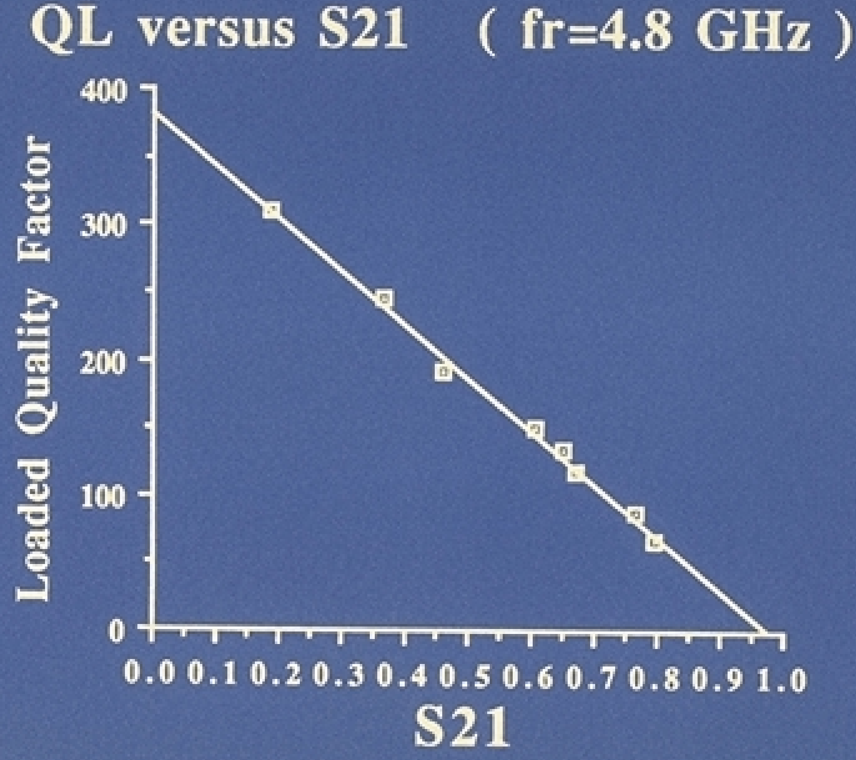


EQUIVALENT CIRCUIT

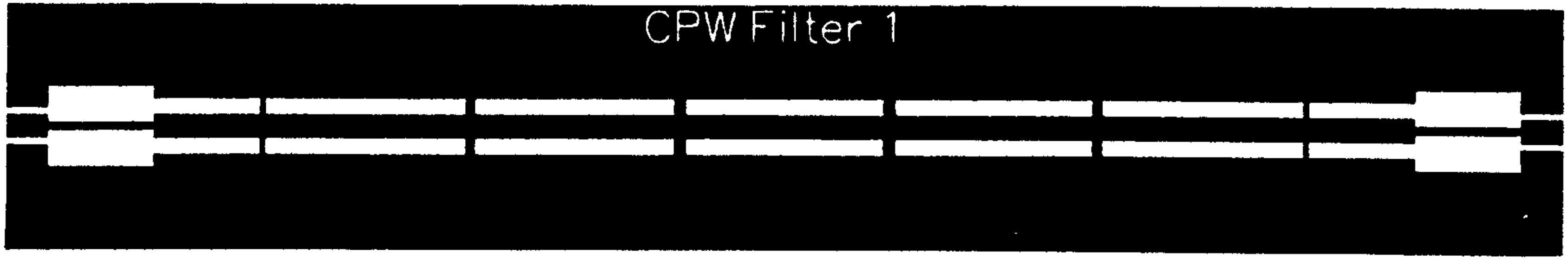
$Q_0 > 500$ at 5GHz on low loss ε_r 2.5 PCB

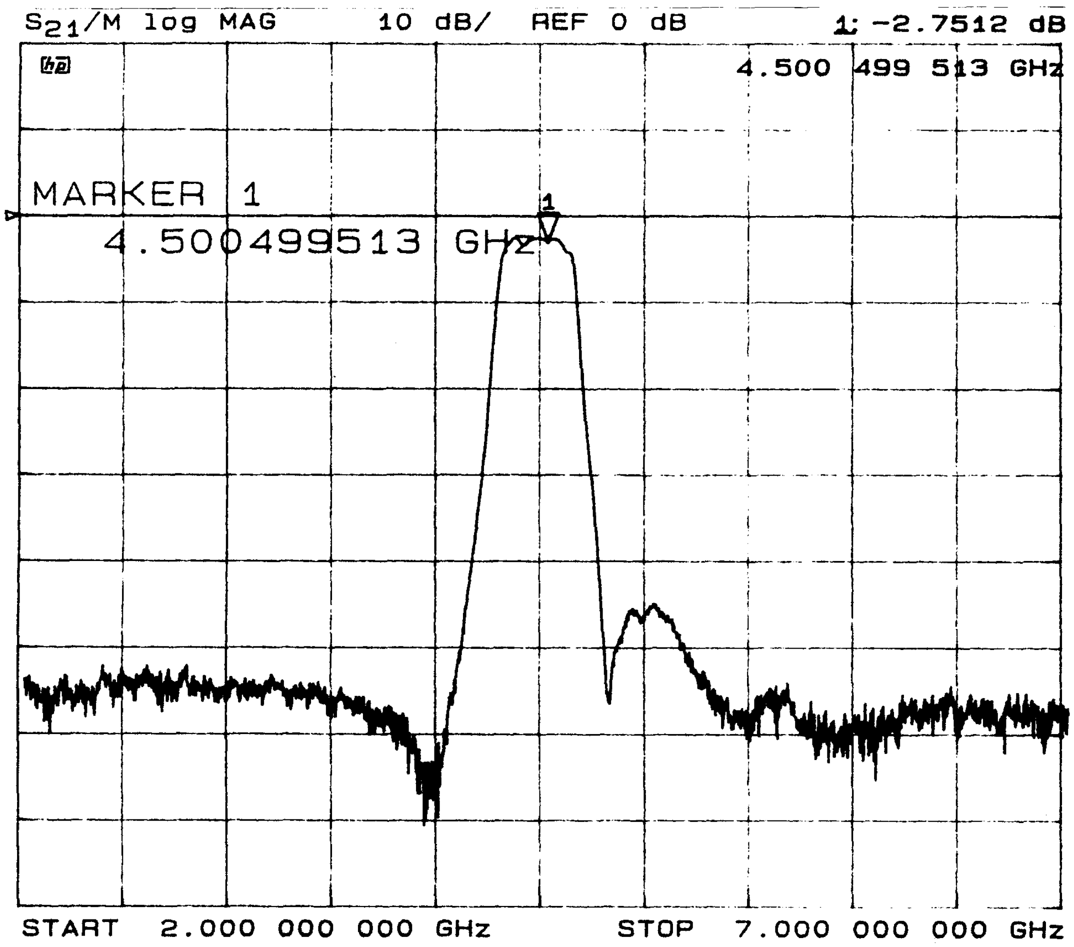
$Q_0 = 380$ at 4.8 GHz on ε_r10 [23]

results without screening, therefore near zero radiation loss



QL versus S21 (fr=4.8 GHz)
Loaded Quality Factor
400
300
200
100
0
0.0
0.1
0.2
0.3
0.4
0.5
0.6
0.7
0.8
0.9
1.0
S21





5 section 4.5GHz bandpass filter

on ε_r10 [23]

low radiation loss

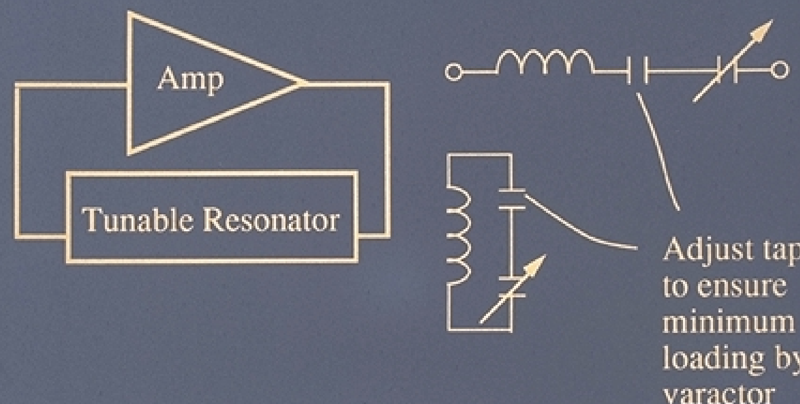
spurious out of band waves exist

Tuning: Varactor Limitations

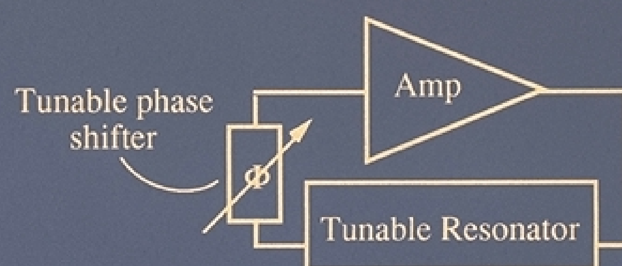
OSCILLATOR TUNING - 2 Main types

1. **Tunable Resonators** Offers broadband and narrowband tuning.

   For low noise reduce loading caused by varactor to minimum



2. **Phase Shift Method** allows narrow band tuning
   Causes $\text{Cos}^4\theta$ degradation



Noise degradation due to varactors Power calculation by - Underhill [25]

- Power dissipated in varactor loss resistor, rs, is: $P = (V_{RS})^2 / rs$
- The voltage across the capacitor in the resonator is: $V_C = QV_{rs}$
- Therefore the power dissipated in the varactor is: $P_V = V_C^{\,2} / Q^2 rs$
- The noise power in oscillators is proportional to $1 / PQ_0^{\,2}$

- The figure of merit $\left(V_C^{\;2}/rs\right)$ should therefore be as high as possible

- Optimum performance obtained for large voltage handling characteristics and small series resistances in varactor

Apply this using new noise equations

- If P is defined as P_{AVO}, $Q_L/Q_0 = 1/2$ and $R_{out} = R_{in}$: then A= 1 and N = 2 then

$$L_{FM} = \frac{2FkT}{Q_0^{\;2} P_{AVO}} \left(\frac{f_0}{\Delta f} \right)^2$$

$$\text{As} \left(L_{FM} = A.\frac{FkT}{8(Q_0)^2 (Q_L/Q_0)^2 (1-Q_L/Q_0)^N P} \left(\frac{f_0}{\Delta f} \right)^2 \right)$$

- As $P_{AVO} = 2P_V$ then

$$L_{FM} = \frac{FkTrs}{V_C^{\;2}} \left(\frac{f_0}{\Delta f} \right)^2$$

Noise performance only dependant on V_C and rs

Example

- A varactor with a series resistance of 1Ω with an RF voltage of 0.25V rms at a frequency of 1GHz.

- The noise performance at 25kHz offset is -97dBc for an amplifier noise figure of 3dB

Improved by

- Reducing the tuning range by coupling varactor into resonator more lightly
- switching in tuning diodes using PIN diodes
- Increased voltage handling using back to back diodes
- improving the varactor

Varactor bias noise

- Flat noise spectral density on bias line causes $(1/\Delta f)^2$ noise in oscillator - same as thermal noise in oscillator -
- For low level modulation:

$$L_{FM} = \frac{(K_F V_M)^2}{(2F_M)^2}$$

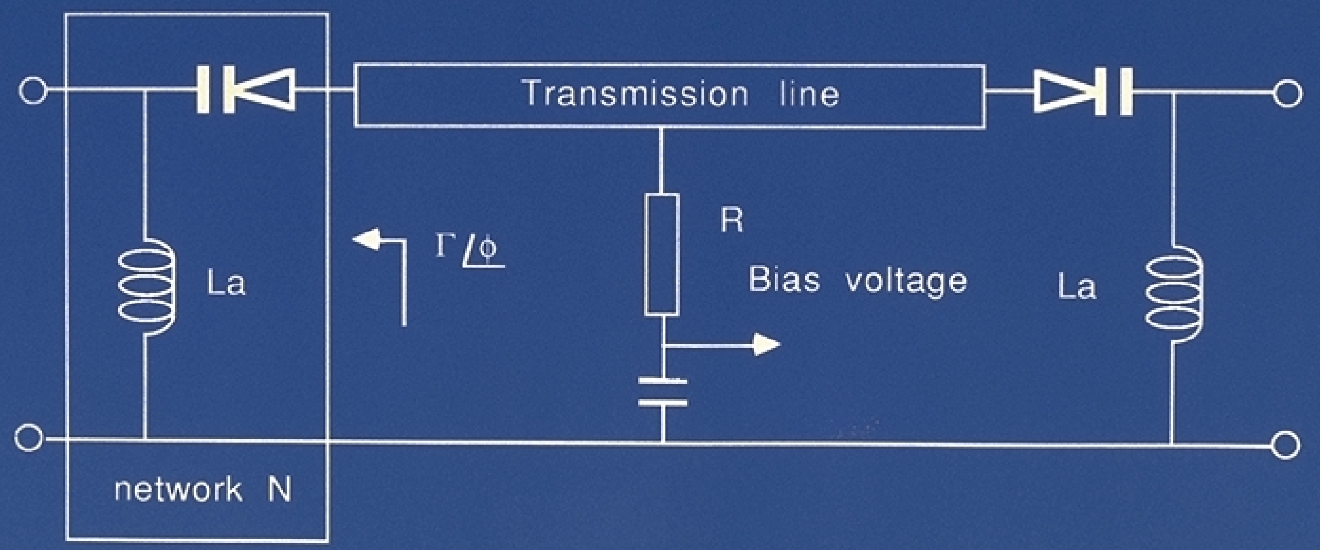
K_F = tuning sensitivity, Hz/Volt
V_M = noise voltage, Volts/√Hz
F_M = offset frequency, Hz

Bias resistor noise

$$\overline{e_n}^2 = 4kTBr_b$$

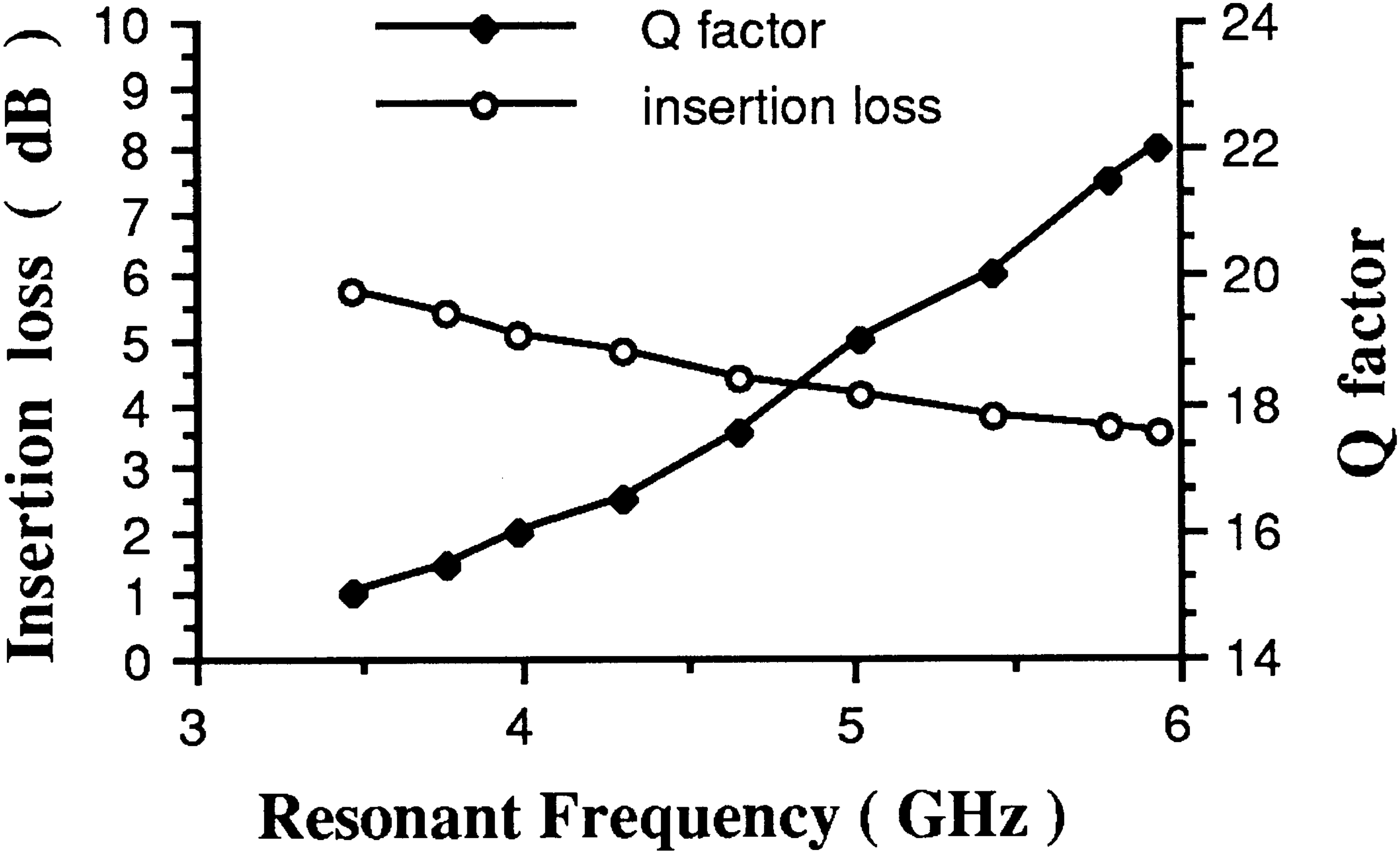
- Keep bias resistor value low
  - less than few hundred ohms eg 50Ω
- Use this to advantage in resonator design to suppress unwanted higher order resonances

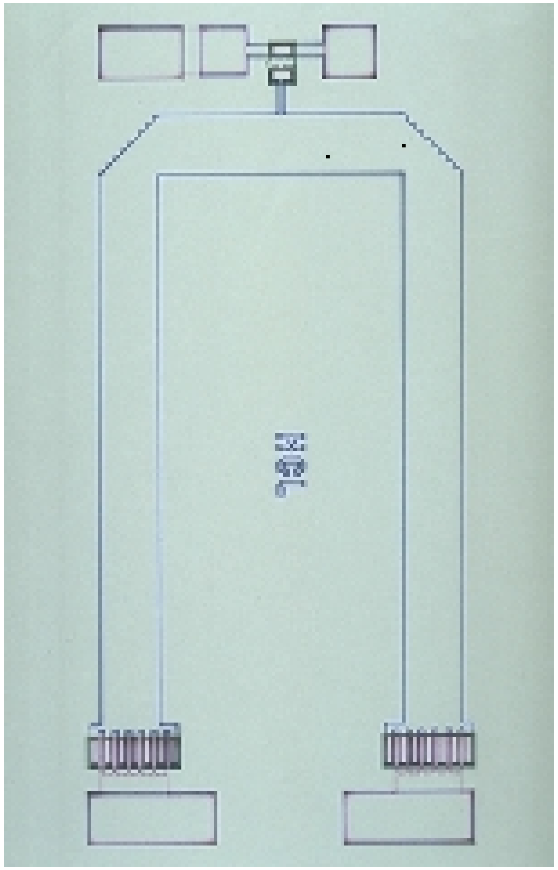
TUNABLE TRANSMISSION LINE RESONATOR



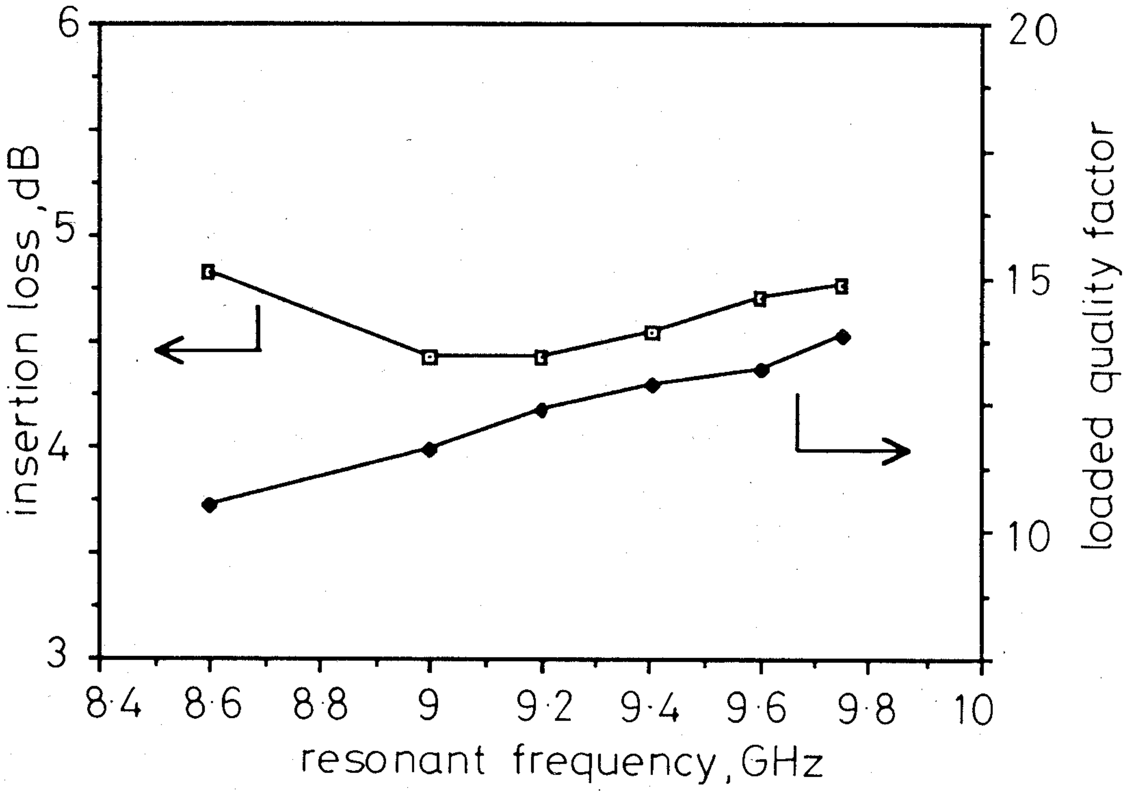
3-6GHz resonator (5mm) on alumina
Two Alpha diodes CVE7900D
C_{j0}=1.5pf, Q (-4V, 50MHz) = 7000
k (capacitance ratio) = 6
[26], [27]

Insertion loss and Q vs frequency, 3-6GHz resonator





2 x 1 mm



8.4 - 9.8GHz GaAs MMIC resonator [27]

Variation of Q_L/Q_0 and S_{21} vs frequency

NOISE DEGRADATION DUE TO OPEN LOOP PHASE ERROR

Effects of open loop phase error [24]

- Always oscillate at N*360º
- Resonator Q degradation as $Q \propto d\phi / d\omega$
- Insertion loss and hence gain increase
- Causes $Cos^4\phi$ degradation in noise performance
- 45º causes 6dB noise degradation
- eg: At 10GHz with DRO Q=10,000, 1MHz offset causes 6dB degradation

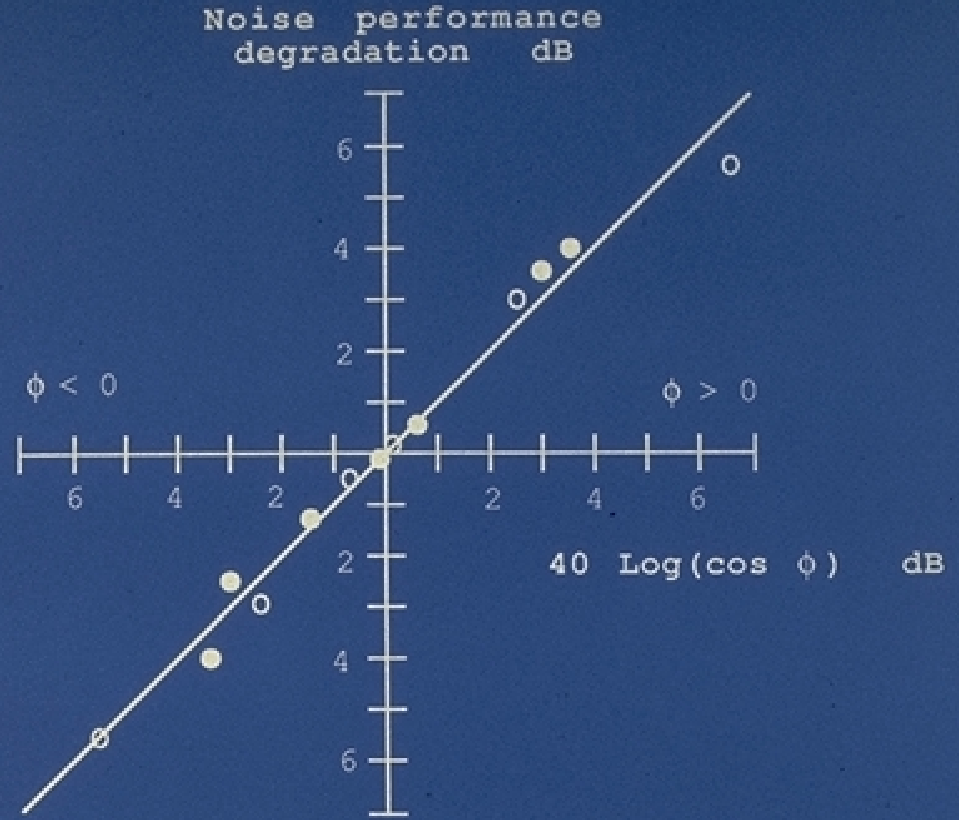


Figure 4 Noise performance degradation with phase error

O - Bipolar ● - GaAs

——— Theory

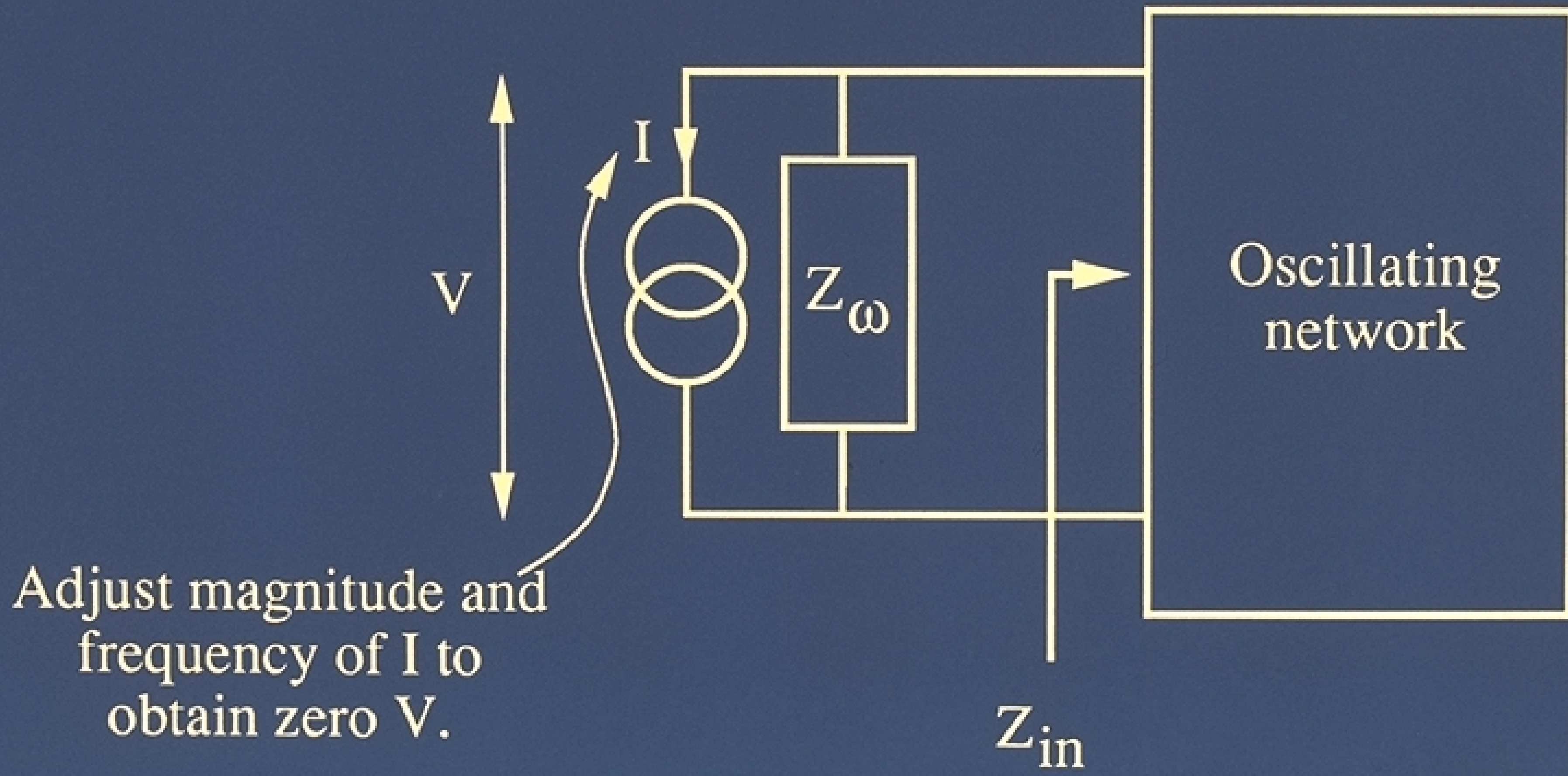
Non linear CAD for oscillators

Non Linear CAD

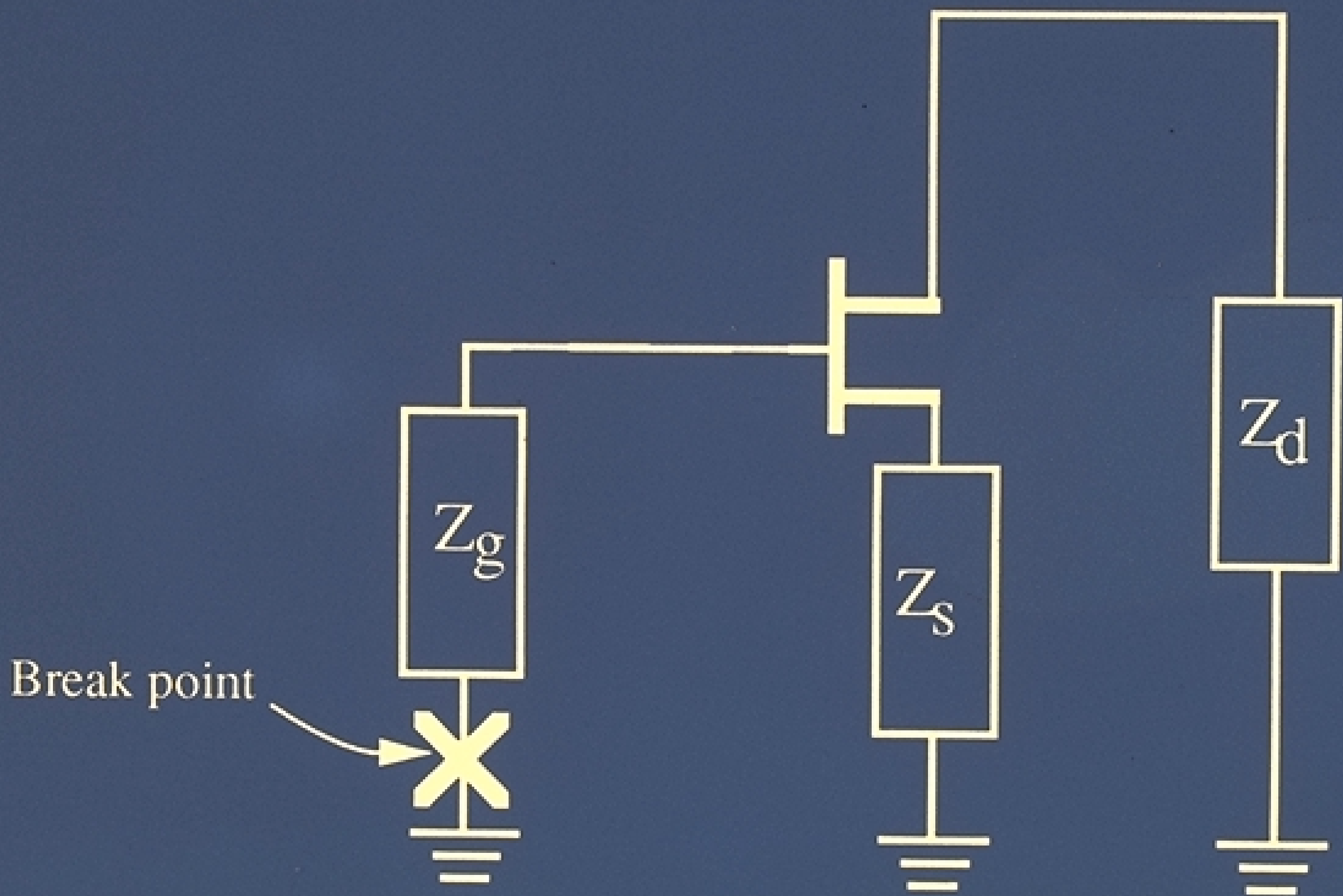
- Break Circuit at short circuit point
- Place current source and frequency dependent resistor at this point
- Make resistor:
  - » Open circuit at fundamental
  - » Short circuit at harmonics
- Adjust amplitude and fundamental frequency of current source to obtain zero volts.

OPTIMISATION TECHNIQUE

Z_ω= O/C at fundamental and S/C at harmonics



TYPICAL OSCILLATOR CIRCUIT



Comparison of computed and measured data

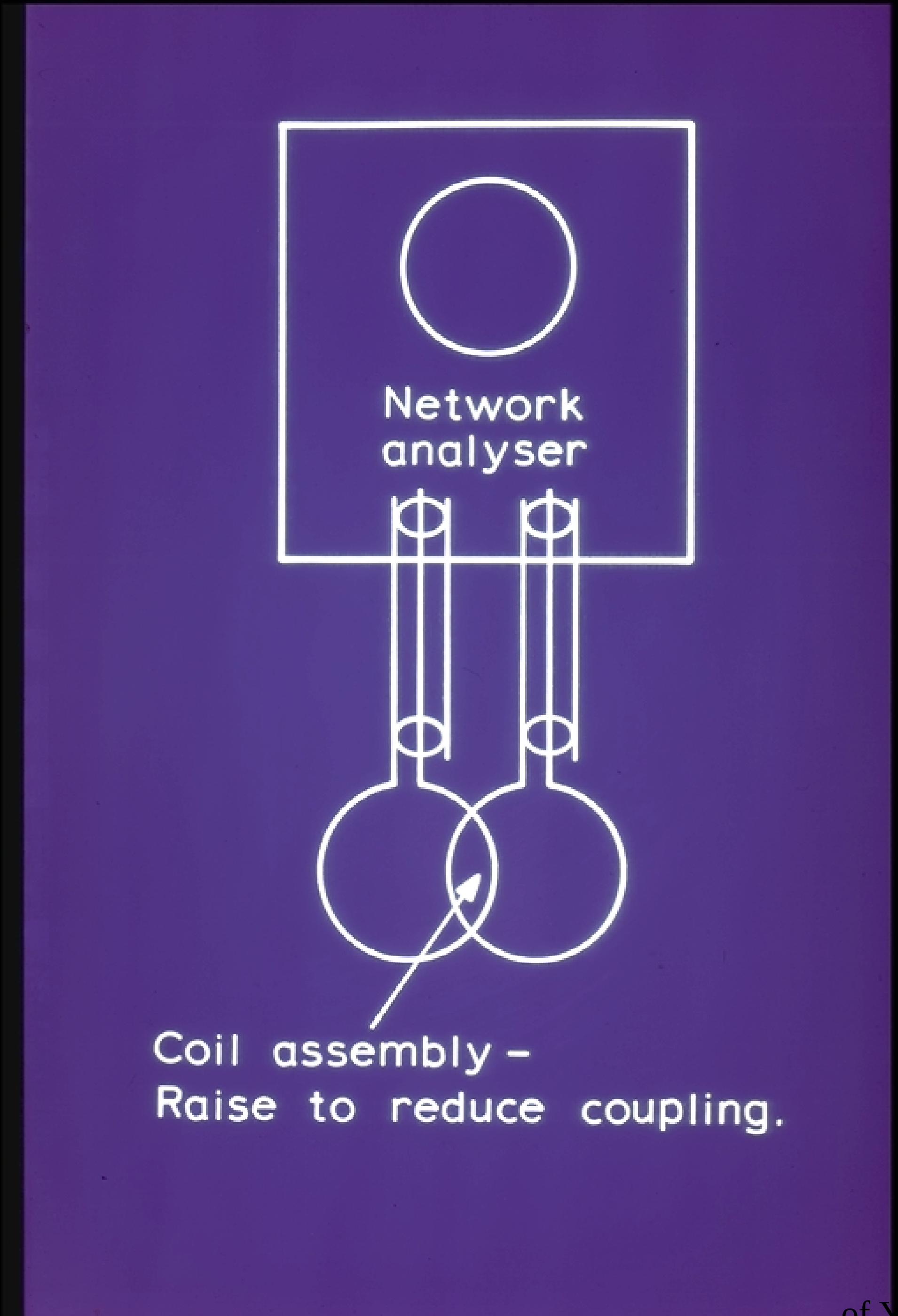
	Predicted	Measured
Resonant Frequency	5.47 GHz	5.41 GHz
Ids	26.5 mA	24.0 mA
Vgs	-0.58 V	-0.53 V
Fundamental	9.3 dBm	8.6 dBm
1 st harmonic	- 14.7 dBm	- 17 dBm
2 nd harmonic	- 20.1 dBm	- 24 dBm

Measurement of coil Q

Adjust coil overlay to obtain low coupling

Place tuned circuit in between coils and measure response

raise to reduce coupling to obtain Q_0



Summary - low phase noise

- High unloaded Q and low noise figure
- Set resonator coupling to achieve $Q_L / Q_0 = 1/2 \rightarrow 2/3$
- Set the open loop phase error to be N.360
- Use a device and circuit configuration producing the lowest transposed flicker noise corner ΔF_C

Summary - low noise tuning

- Incorporate varactor loss resistor into resonator and set Q_L/Q_0 as before
- For narrow band tuning
  - loosely couple varactor into resonator and set Q_L/Q_0 as before or consider
  - low loss phase shifter in the feedback loop. Expect 6dB noise degradation if open loop phase error goes to 45 degrees
- Arrange for low bias line noise eg $r_b=50\Omega$

LC Design Example

Design Example

Design a 150 MHz oscillator using:

1. 235nH inductor with a Q_0 of 300.

2. An inverting amplifier with an input and output impedance of 50Ω

An LC resonator with losses can be represented as an LCR resonator as shown in Figure 24.

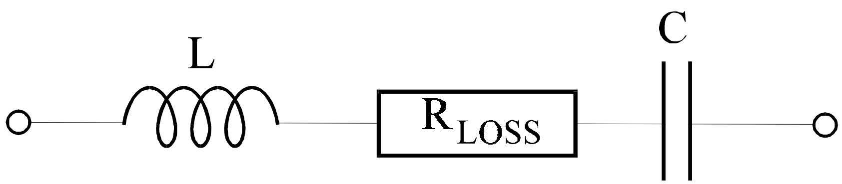


Figure 24 Model of LC resonator including losses

As $$Q_0 = \frac{\omega L}{R_{loss}}$$

The equivalent series resistance is 0.74Ω.

Assuming the amplifier has an O/P impedance R_{out}, The ratio Q_L/Q_0 is:

$$\frac{Q_L}{Q_o} = \frac{R_{loss}}{R_{loss} + R_{in} + R_{out}}$$

For $R_{in} = R_{out}$

$$\frac{Q_L}{Q_o} = \frac{R_{loss}}{R_{loss} + 2R_{in}}$$

Let $\frac{Q_L}{Q_o} = \frac{1}{2}$ as the P_{AVO} definition can be used.

then $R_{loss} = 2\,R_{in}$ then $R_{in} = \frac{R_{loss}}{2}$

Therefore $R_{in} = R_{out} = 0.37\Omega$ as shown in Figure 25.

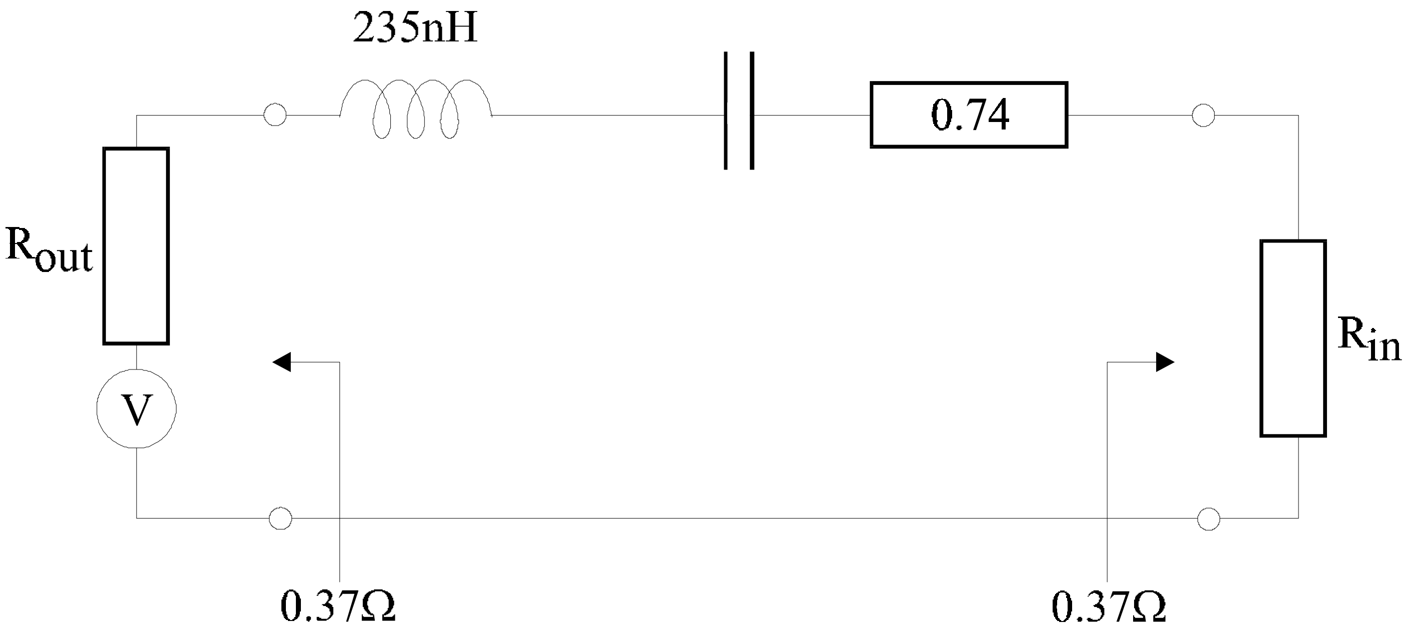


Figure 25 LC resonator with scaled source and load impedances

As amplifier has 50Ω input and output impedances:

Use LC transformer as shown in Figure 26.

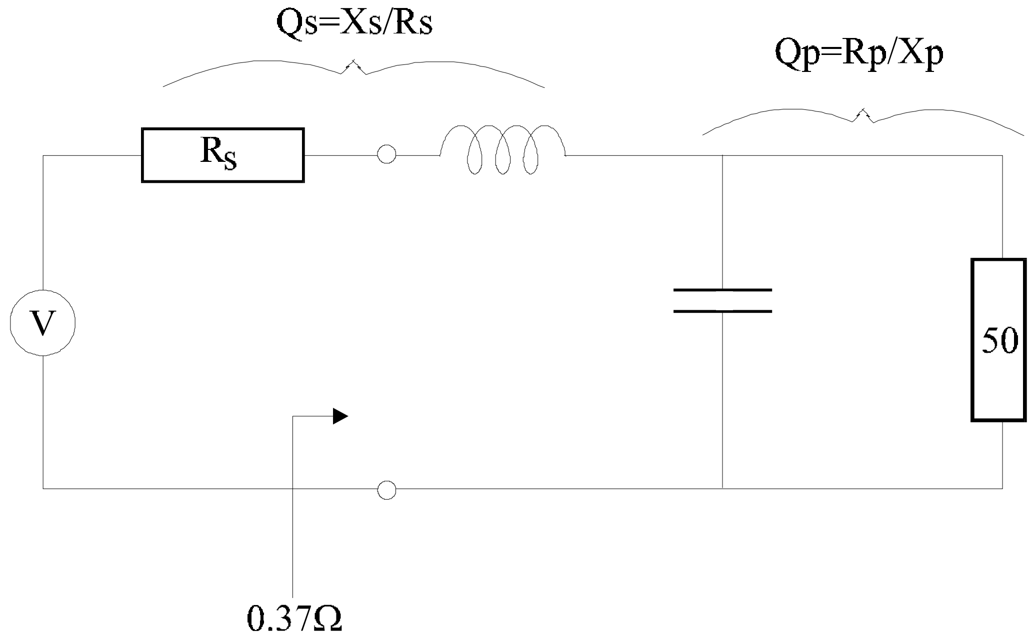


Figure 26 LC transformer to convert to 50Ω

The equations for the series and shunt components are:

$$Q_s = Q_p = \sqrt{\left(\frac{R_p}{R_s} - 1\right)}$$

The Q of the series component is:

$$Q_S = \frac{X_s}{R_s}$$

The Q of the shunt component is:

$$Q_p = \frac{R_p}{X_p}$$

Note:

R_p = shunt resistance

R_S = series resistance

X_S = series reactance = $j\omega L$

X_p = shunt reactance = $\frac{1}{j\omega c}$

$$Q_s = Q_p = \sqrt{\left(\frac{R_p}{R_s} - 1\right)} = \sqrt{\left(\frac{50}{0.37} - 1\right)} = 11.58$$

$X_s = 4.28 = j\omega L$ $L = 4.5$ nH

$X_p = 4.31 = \dfrac{1}{j\omega C}$ $C = 246$ pf

Incorporate 2 transforming circuits into resonator circuit as shown in Figure 27

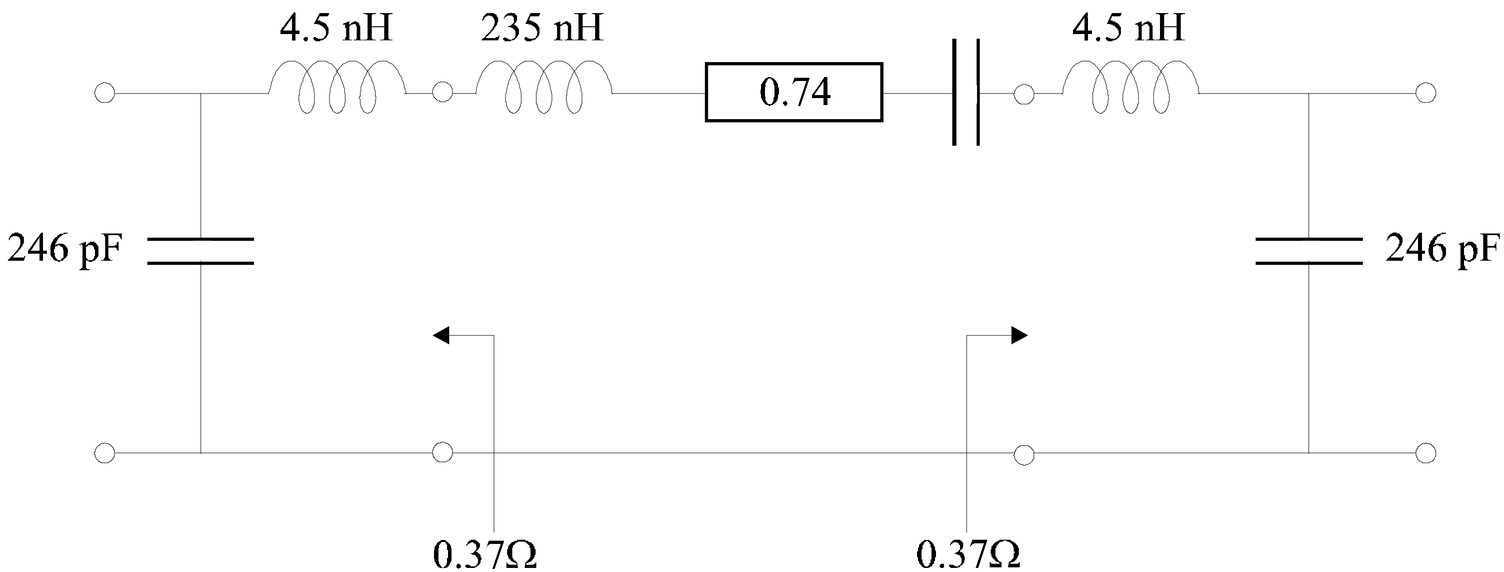


Figure 27 LC resonator with impedance transformers

As the total inductance is 235nH, the part that resonates with the series capacitor is reduced by 9nH as shown in Figure 28.

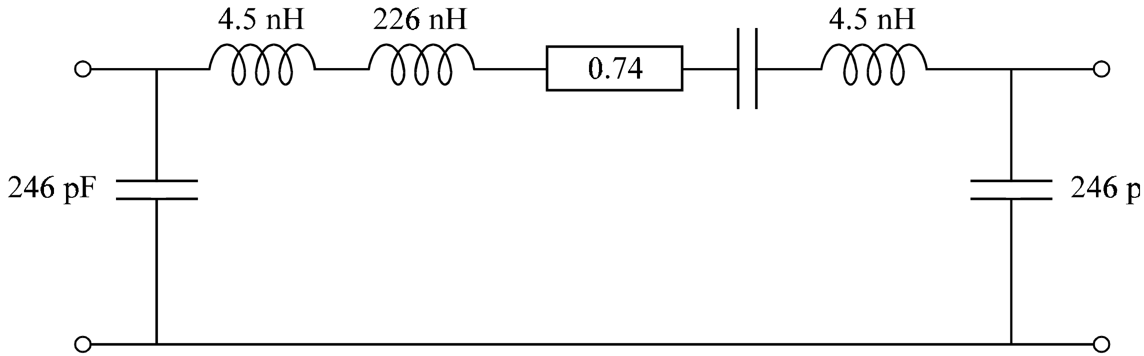


Figure 28 Resonator with total L = 235nH

It is now necessary to calculate the resonant frequency.

The part of the inductance which resonates with the series capacitance is reduced by the matching inductors to:

235nH - (2 x 4.5 nH) = 226 nH

$$f = \frac{1}{2\pi\sqrt{LC}} \qquad LC = \frac{1}{(2\pi f)^2}$$

$$C = \frac{1}{L(2\pi f)^2} = 5pf$$

Note the value of shunt capacitors: 246pf!

The circuit now becomes:

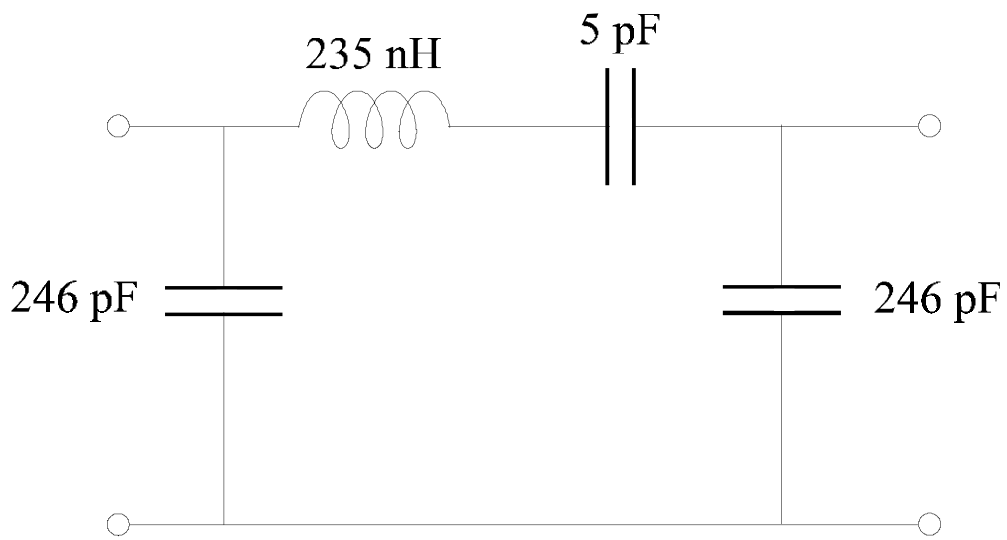
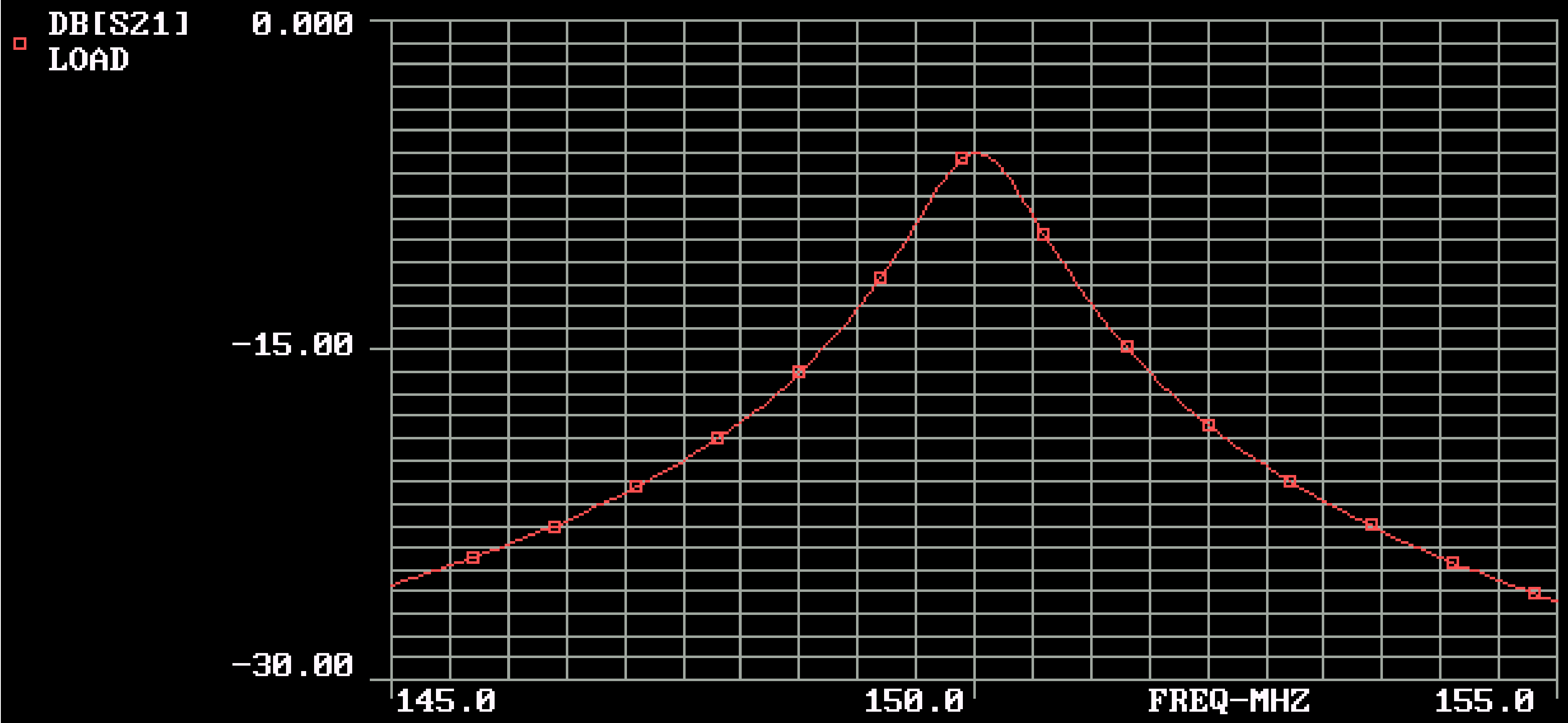


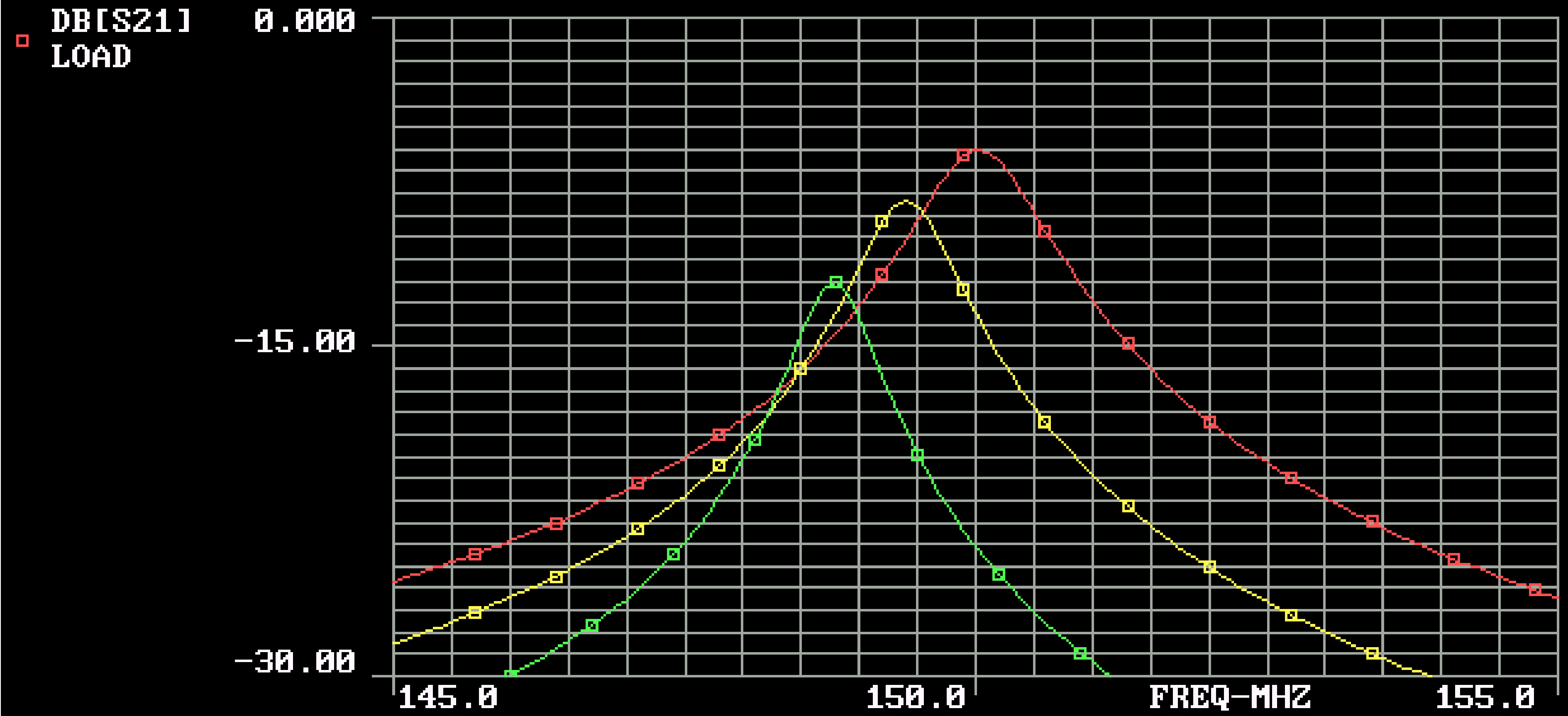
Figure 29 Final resonator circuit



Simulation of insertion loss, S_{21}, of resonator

Effect of parasitic components

- What is the effect of the parasitics in the shunt capacitors
- Investigate the effect of both a 1nH and 2nH parasitic inductance
- This increases the effective capacitance as close to resonance (reduces impedance)



Effect of parasitic inductance in shunt C

Yellow = 1nH

Green = 2nH - correct for this by reducing C from 246pf to 174pf

Note the phase shift at resonance is 180^o.

So the amplifier should provide a further 180^o.

If necessary a phase shifter should be included to ensure N x 360^0 at the peak in the resonance as shown in Figure 30.

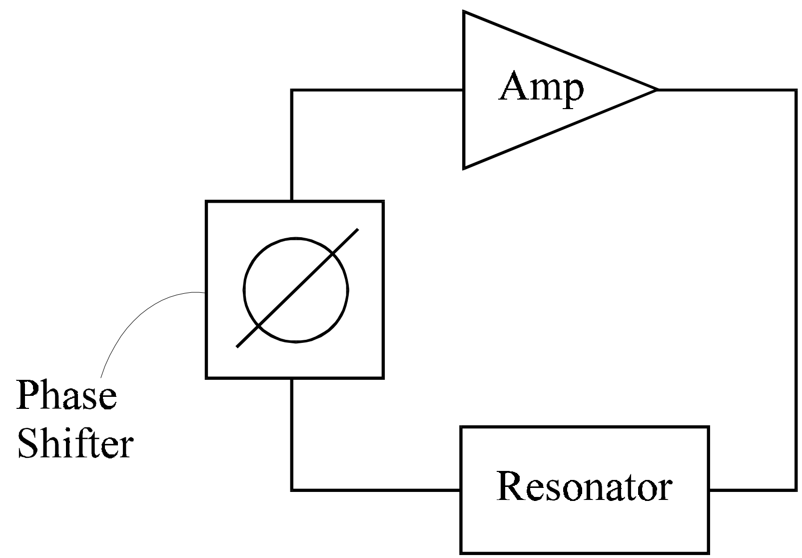
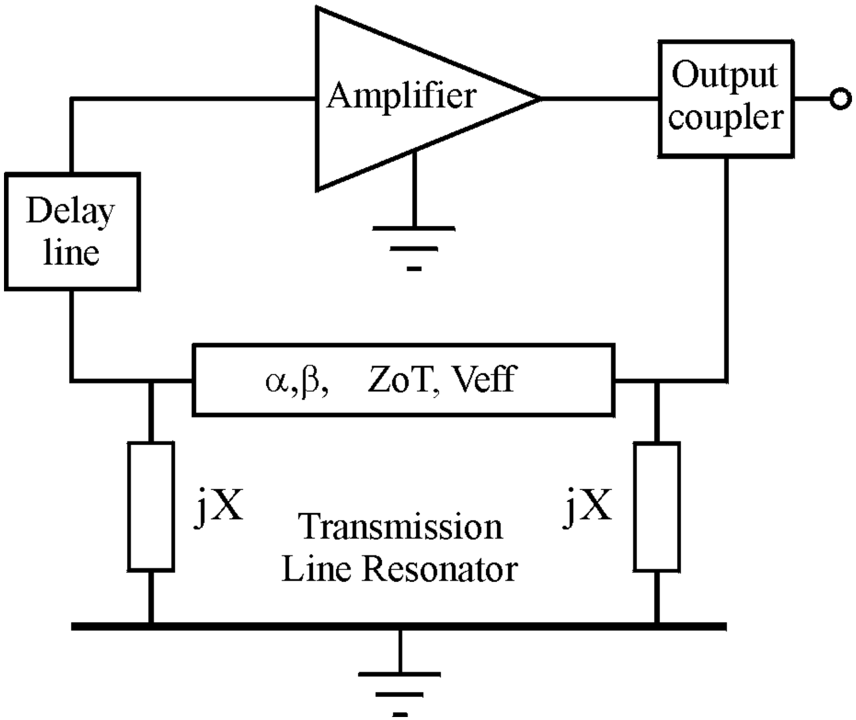


Figure 30 Oscillator incorporating phase shifter

1 GHz Transmission Line Osc.

- Design a 1GHz Transmission line Oscillator use 1.5mm FR4 PCB, $Z_0 = 50\Omega$, $\varepsilon_{eff} = 3.3$



Assume $Z_0T = Z_0line$

$$Q_0 = \frac{\pi}{2\alpha L} \qquad Q_L = \pi S_{21}(0)\left(\frac{X^2}{4}\right)$$

$$Length\ L = \left(\frac{V_{eff}}{2 f_0}\right)\left(1 + \left(\frac{1}{\pi}\right)\tan^{-1}\left(\frac{2}{X}\right)\right)$$

$X = 2\pi f_0 C Z_0$ or $-Z_0/2\pi f_0 l$

Find loss of line

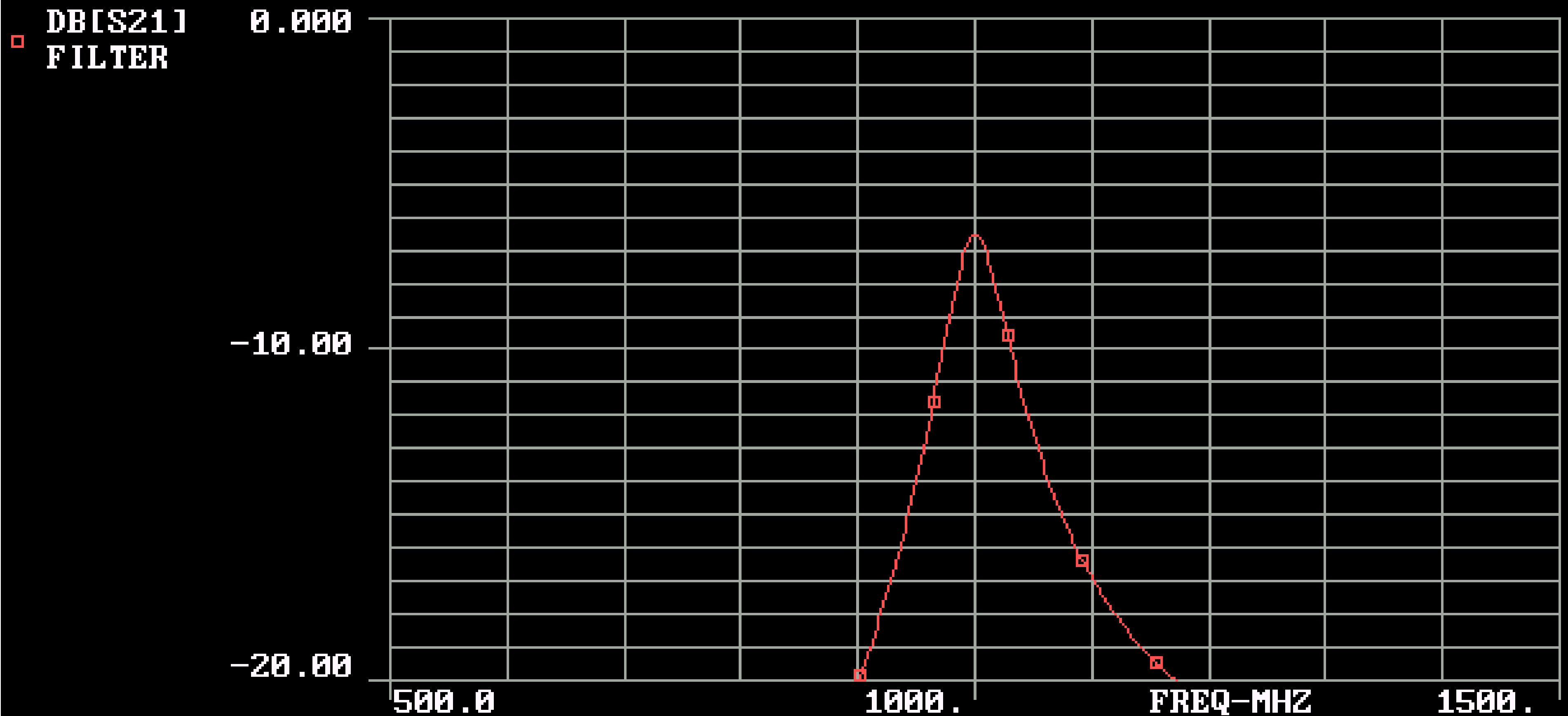
- Measure loss of known length of line OR
- Build a number of resonators with varying Q_L. Extrapolate to Q_0 by drawing straight line as $S_{21} = \left(1 - Q_L / Q_0\right)$
- Simulate using field model
- Note that the loss of 'low loss' transmission lines can be deduced from resonator measurements

Calculate parameters

- From measurement $Q_0 = 39.3$
- $\alpha L = 0.04$
- For $Q_L/Q_0 = 1/2$ $\qquad X = \sqrt{\dfrac{2}{\alpha L}}$
- $X = 7.07$ therefore as $X=2\pi f_0 C Z_0 = -Z_0/2\pi f_0 l$
  - inductor l = 1.125nH
  - capacitor C = 22.5pf

$$L = \left(\frac{V_{eff}}{2 f_0}\right)\left(1+\left(\frac{1}{\pi}\right)\tan^{-1}\sqrt{2\alpha L}\right)$$

- Line length=7.53cms for l and 8.98cms for C



Transmission line resonator response using shunt inductor

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