第二章 掃瞄式頻譜分析儀簡介

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2.5 量測實例與其它量測配置

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感謝安捷倫公司同意使用安捷倫發展的教材

2.1 訊號分析

- 訊號可以在時域檢視(e.g., 使用示波器)，也可以在頻域檢視(e.g., 使用頻譜分析儀)。
- 傳統類比示波器可以觀察週期性訊號的振幅、週期與頻率，數位儲存式示波器可進一步觀察非週期性訊號中任何一段時間之內的訊號波形。
- 傳統頻譜分析儀可以觀察訊號的頻率、功率、強度或功率在頻率軸上的密度、諧波、雜訊等等，但無法觀察訊號的相位；若搭配追蹤訊號源(tracking generator)，則可進一步量測兩端口元件的穿透的振幅頻率響應，包括元件的非線性特性。

- 時域的訊號表達式與頻域的訊號表達式之間具有一對一的關係，如下列的 **Fourier-Transform pair**

$$G(f)=\int_{-\infty}^{\infty} g(t)e^{-j2\pi ft}df = F\{g(t)\}$$
$$g(t)=\int_{-\infty}^{\infty} G(f)e^{j2\pi ft}dt = F^{-1}\{G(f)\}$$

- 根據 **Fourier Series** 的理論，何一個週期性訊號皆可表示為

$$g(t)=a_0+2\sum_{n=1}^{\infty}\left[a_n\cos(2\pi nf_0t)+b_n\sin(2\pi nf_0t)\right]=\sum_{n=-\infty}^{\infty}c_ne^{j2\pi nf_0t}$$

其中

$$c_n=\frac{1}{T_0}\int_0^{T_0}g(t)e^{-j2\pi nf_0t}dt$$

- 若把週期性脈波表示為

$$g(t)=\sum_{n=-\infty}^{\infty}g_{T_0}(t-nT_0)$$

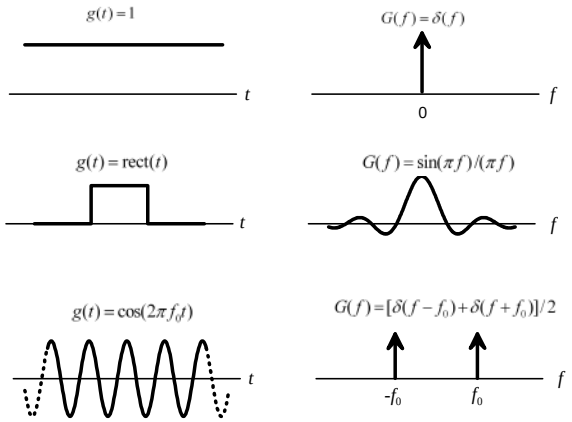
其中

$$g_{T_0}(t)=\begin{cases}g(t), & 0\le t<T_0\\ 0, & \text{otherwise}\end{cases}$$

是 ***g(t)*** 的基本脈波。則

$$c_n=\frac{1}{T_0}\int_0^{T_0}g(t)e^{-j2\pi nf_0t}dt=\frac{1}{T_0}\int_0^{T_0}g_{T_0}(t)e^{-j2\pi nf_0t}dt$$
$$=\frac{1}{T_0}\int_0^{\infty}g_{T_0}(t)e^{-j2\pi nf_0t}dt=\frac{1}{T_0}G_{T_0}(nf_0)=\frac{1}{T_0}G_{T_0}(f)\Big|_{f=nf_0}$$

這個式子提供了**Fourier Series** 的係數與 **Fourier Transform** 的連結



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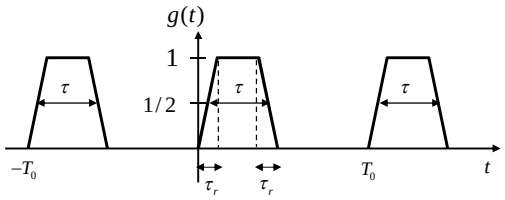

一個週期為 $T_0=1/f_0$ 的梯形脈波序列， 責任週期為 $d = \tau/T_0$ ，上升(下降)速率為 $d_r = \tau_r/T_0$ 。 若 $d_r = 0$ ，則簡化為矩形脈波序列。

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- 若為矩形脈波序列，其基本矩形脈波可表示為

$$g_{T_0}(t) = \text{rect}\left(\frac{t-(\tau/2)}{\tau}\right), \quad \text{with} \quad \text{rect}(t) = \begin{cases} 1, & -\frac{1}{2} \le t < \frac{1}{2} \\ 0, & \text{otherwise} \end{cases}$$

因此

$$c_n = \frac{1}{T_0} G_{T_0}(nf_0) = \frac{1}{T_0} F\left\{\text{rect}\left(\frac{t-(\tau/2)}{\tau}\right)\right\}\Bigg|_{f=nf_0} = d\,\frac{\sin(\pi nd)}{\pi nd} e^{-j\pi nd}$$

頻譜上限 (upper bound) 為

$$\frac{1}{T_0}\left|G_{T_0}(f)\right|_{\text{bound}} \approx \begin{cases} d, & f \to 0 \\ 1/(\pi f / f_0), & f/f_0 >> 1/(\pi d) \end{cases}$$

是 0 dB/decade 與 -20 dB/decade 的兩條漸近線

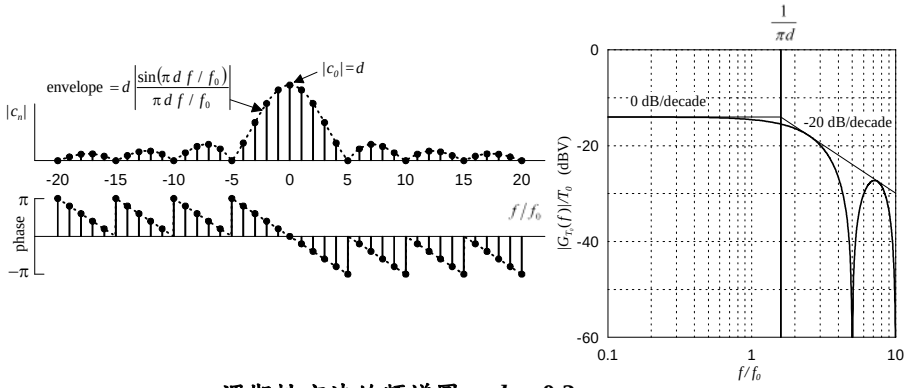
- 若為梯形脈波序列，則

$$c_n = d\, e^{-j\pi nd} e^{-j\pi nd_r} \frac{\sin(\pi nd)}{\pi nd} \frac{\sin(\pi nd_r)}{\pi nd_r}$$

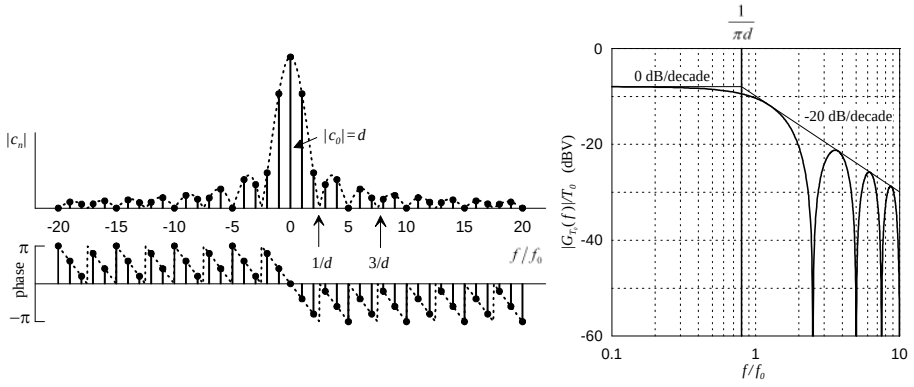
其頻譜上限 (upper bound) 為

$$\frac{1}{T_0}\left|G_{T_0}(f)\right|_{\text{bound}} = \begin{cases} d, & f/f_0 \le \frac{1}{\pi d} \\ \frac{1}{\pi f / f_0}, & \text{if } \frac{1}{\pi d} \le f/f_0 \le \frac{1}{\pi d_r} \\ \frac{1}{(\pi f / f_0)^2 d_r}, & f/f_0 \ge \frac{1}{\pi d_r} \end{cases}$$

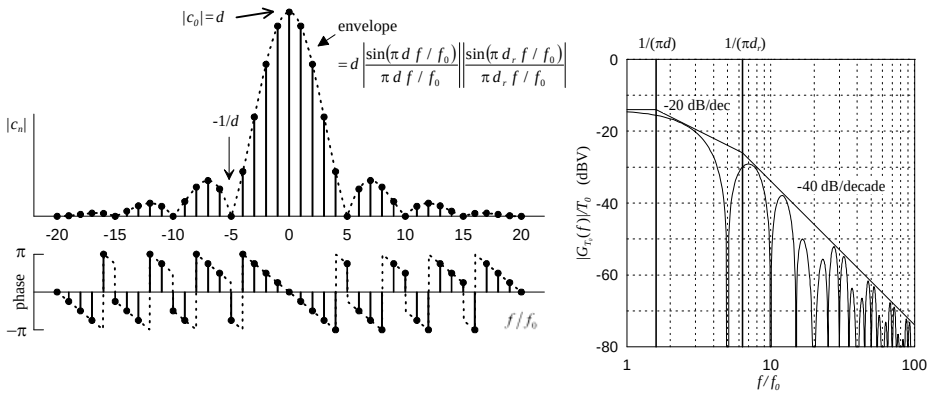
是 0 dB/decade 、 -20 dB/decade 與 -40 dB/decade 的三條漸近線



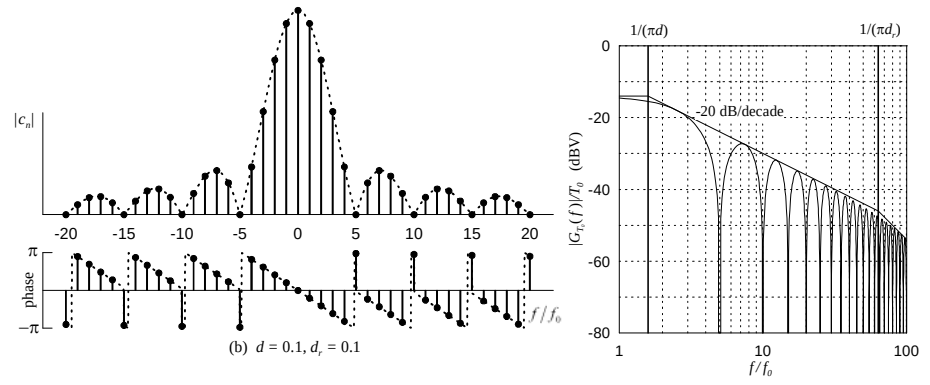
週期性方波的頻譜圖，d = 0.2。



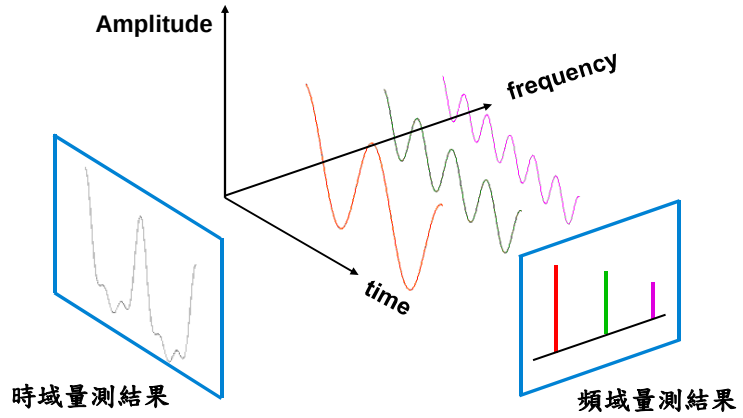
週期性方波的頻譜圖，d = 0.4。



週期性梯形波的頻譜圖，d = 0.2，d_r = 0.05。



週期性梯形波的頻譜圖，d = 0.2，d_r = 0.005。



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2.2 頻譜分析儀的種類

- 濾波器群 (Filter-Bank) SA
- 濾波器掃瞄式 (Swept-Filter) SA
- 本地振盪器掃瞄式 (Swept-LO) SA 或掃瞄式 SA
- 快速富立葉變換 (FFT) SA

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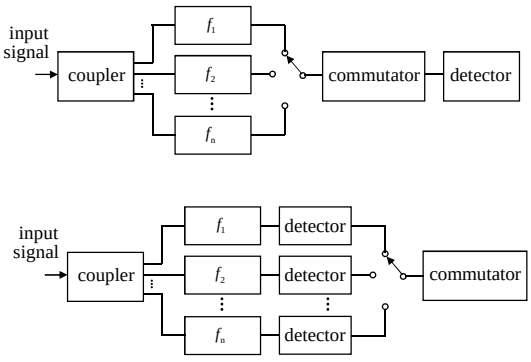
富立葉頻譜分析儀（快速富立葉變換 SA）

- 由於受到 IC 速度的限制，傳統的富立葉頻譜分析儀只適用於低頻訊號。
- 近年來，IC 速度大幅提升(包含取樣與 A/D 轉換)，加上數位訊號處理的進步，此 SA 可以分析數位調變的向量訊號，功能強大，非傳統 SA 可匹敵，因此，已經成為高級頻譜分析儀的代名詞。

頻譜分析儀的種類

- **Filter-bank SA**



由一群 n 個頻率相鄰的率波器組成，輸入訊號同時進入到這 n 個頻寬為 B 的濾波器，這些濾波器的輸出端共用或個別使用一個紀錄器。

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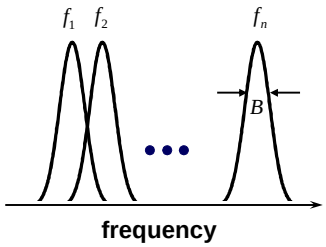
Filter-bank SA



- 個別濾波器的頻寬越寬，解析度越高，但對相同的頻率量測範圍而言，所需的濾波器個數越多。
- 概念簡單，但每個濾波器的中心頻率皆不同，製造困難，成本高。

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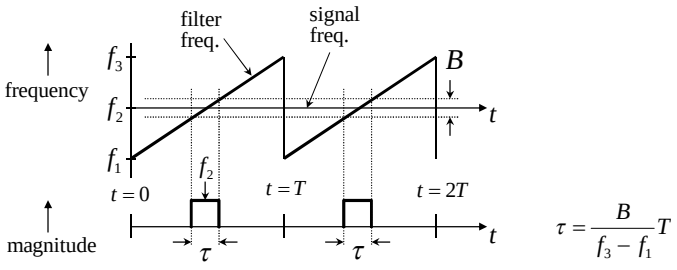
濾波器掃瞄式 (Swept-Filter) SA



- 輸入訊號的頻率為 f_2，具有頻寬為 B 的濾波器之中心頻率可從 f_1 隨時間增加調整到 f_3，在濾波器的中心頻率靠近 f_2 時，濾波器輸出一個脈波，此脈波的時間寬度為 τ。
- 此種可快速控制中心頻率的濾波器複雜度與製作難度都高。

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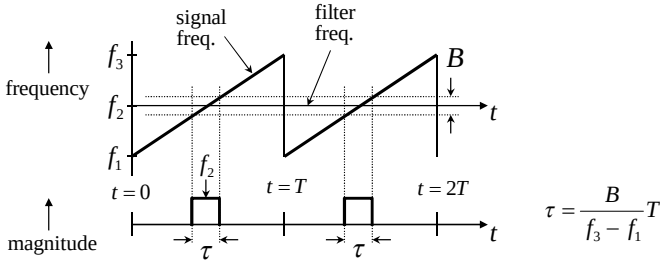
改變輸入訊號的頻率



- 若具有頻寬 B 的濾波器之中心頻率固定在 f_2，而輸入訊號的頻率被從 f_1 隨時間增加調整到 f_3，則濾波器會有與上一張投影片相同的輸出(重要的是濾波器之中心頻率與輸入訊號頻率的相對關係，誰被固定或誰被調整並不重要)。
- 輸入訊號中心頻率的改變可以透過混頻器來達成。
- 這構成本地振盪器掃瞄式 SA 的基礎。

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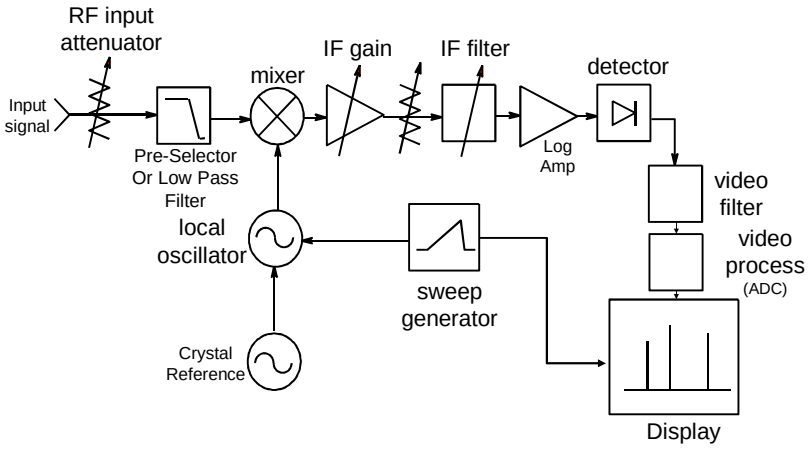
2.3 掃瞄式頻譜分析儀之方塊圖



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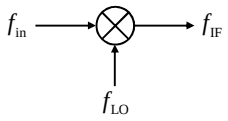
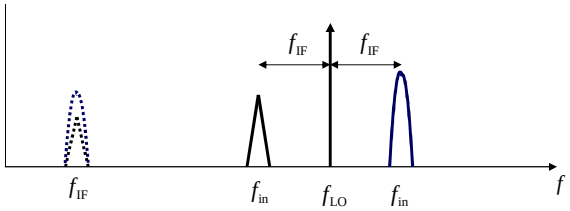
混頻器(Mixer)



$$\cos(2\pi f_{LO}t)\cos(2\pi f_{in}t)$$
$$=\frac{1}{2}\{\cos[2\pi(f_{LO}+f_{in})t]+\cos[2\pi(f_{LO}-f_{in})t]\}$$
$$f_{IF}=|f_{LO}\pm f_{in}| \qquad f_{in}=|f_{LO}\pm f_{IF}|$$



在 input 前面必須加一個濾波器，以便選擇在互為 image frequency 位置的兩個訊號的其中一個。

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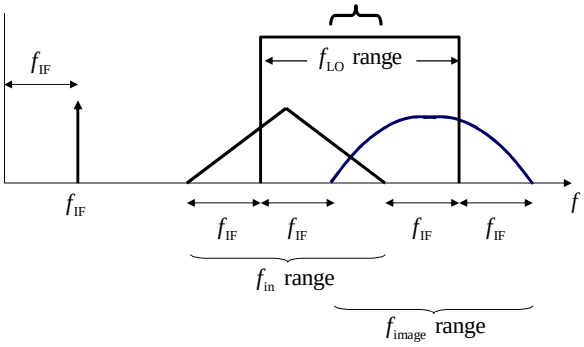
f_{IF} 的選擇($f_{IF} < f_{in,min}$)

overlap of f_{in} and f_{image} range
if $(f_{in,max} - f_{in,min}) > 2f_{IF}$



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f_{IF} 的選擇($f_{IF} < f_{in,min}$)

- 若 $f_{IF} < f_{in,min}$，且 $(f_{in,max} - f_{in,min}) > 2 f_{IF}$ ，就會發生 input 頻率範圍與 image 頻率範圍互相重疊的情況，使得這兩種訊號無法被固定的濾波器分離。
- 若要避免 image 訊號被轉換到 f_{IF} 頻帶，就必須使用追蹤帶通濾波器(tracking BPF) ，此 BPF 只截取 input signal 部分頻率位置的成份，因為 f_{LO} 改變後，此 BPF所截取的 input signal 頻率成份也必須跟著改變，因此其中心頻率必須隨著 f_{LO} 的改變而改變。
- 此 BPF 複雜且昂貴。
- Input signal 頻率超過 3 GHz 的 SA 通常是採取這種 $f_{IF} < f_{in,min}$ 的架構。

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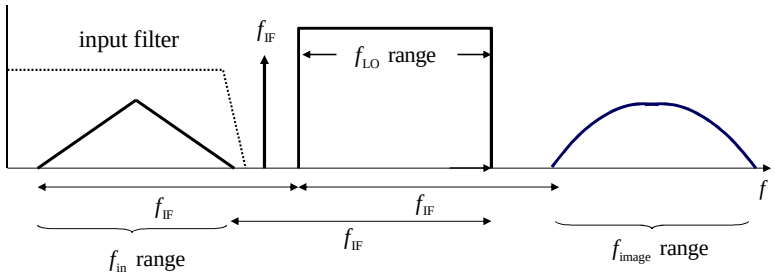
f_{IF} 的選擇($f_{IF} > f_{in,max}$)



- 若 $f_{IF} > f_{in,max}$，就能避免 input 頻率範圍與 image 頻率範圍互相重疊的情況，這兩種訊號可藉由固定的低通濾波器分離。
- 此濾波器成本較低廉， Input signal 頻率低於 3 GHz 的 SA 通常是採取這種 $f_{IF} > f_{in,max}$ 的架構 。

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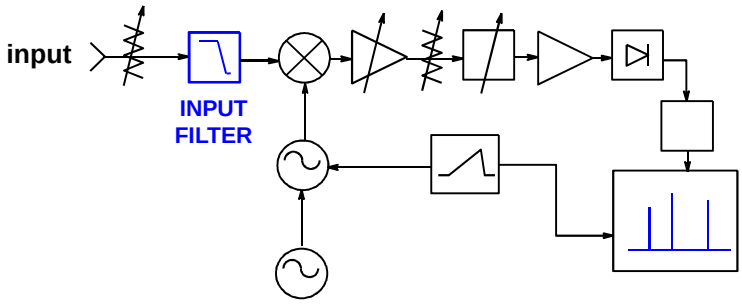
混頻器前端低通濾波器



- 這種 $f_{IF} < f_{in,min}$ (低中頻)的架構，則為中心頻率被 LO 頻率追蹤的 BPF。
- 若是 $f_{IF} > f_{in,max}$ (高中頻)的架構，則為固定頻寬的 LPF 。

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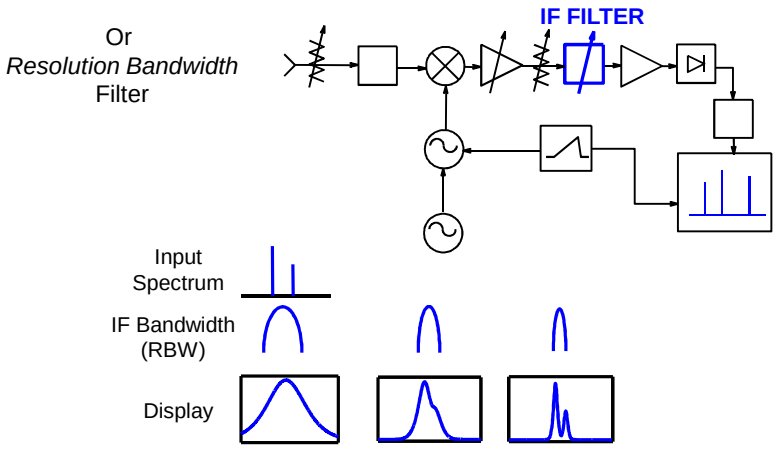
中頻濾波器(解析頻寬濾波器)



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中頻濾波器(解析頻寬濾波器)

- RBW 的大小選擇不同，可能導致不同的量測結果，What you see may not be what you have 。
- 較窄的 RBW ，選擇性較佳，換言之，解析度較好(比較能分辨兩個頻率靠很近的單頻訊號) 。
- 較窄的 RBW 使得通過 IF BPF 而被 detect 的雜訊減少 (SNR 增加)，使得較弱的 input signal 成份被般淹沒的機會減少，因此可提高量測的動態範圍。
- 較窄的 RBW 必須配合較長的掃瞄時間(後續討論) ，影響量測時間與效率。
- 就 BPF 裙擺選擇性的觀點，brick-wall 型態的 IF BPF 是最理響的選擇，然而這種 BPF 的暫態時間太長，所以 Gaussian 型態的 IF BPF 是最務實的。

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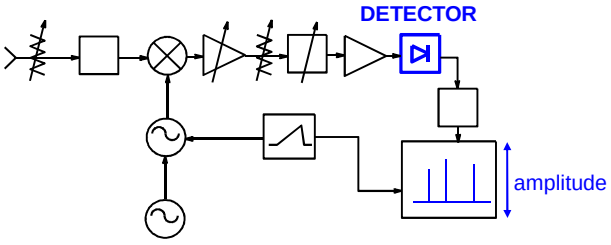
檢測器



- 檢測器(或是包絡檢測器)可以將通過 IF filter 之後的 modulated signal 的 envelope 萃取出來，經由 ADC 再將資料點顯示在螢上。
- 如果顯示的頻率範圍很大 ，或是 RBW 很小(注意： LO 的 frequency step < 0.1 RBW)，導致螢幕的每一個 pixel 要表示很多頻率點的資料。
- 不同的 detection mode 會顯示不同的結果。

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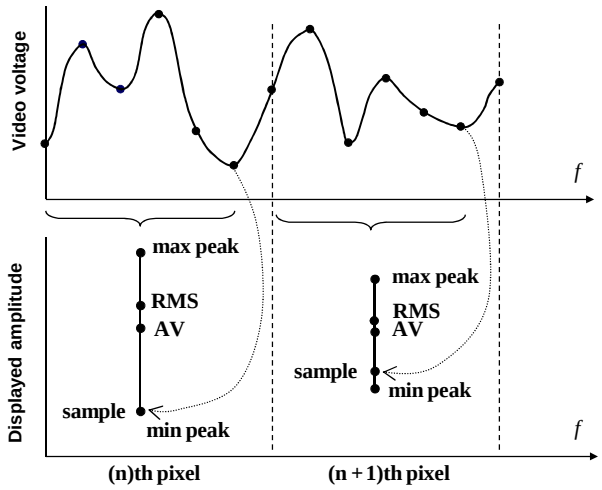
檢測器之 detection mode



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本頁取材自 Rauscher

檢測器之 detection mode

- **Max Peak (positive Peak) detector： 每一個 pixel 包含的所有 samples 中的最大值 (good for analyzing sinusoids, but tends to over-respond to noise when so sinusoids are present)**
- **Min Peak (negative Peak) detector： 每一個 pixel 包含的所有 samples 中的最小值**
- **Sample detector ：每一個 pixel 包含的所有 samples 中的最後一個取樣值(best for noise-like signals; may miss the peaks of burst and narrowband signals when RBW is smaller than the frequency spacing for adjacent pixels)**
- **RMS detector ：** $V_{rms} = \frac{1}{M}\sum_{i=1}^{M} v_i^2$
- **AV (average) detector ：** $V_{avg} = \frac{1}{M}\sum_{i=1}^{M} v_i$
- **Normal detector (Rosenfell detector) ：對 CW-like signal 顯示 pos peak ，對 noise 交替顯示 pos peak 與 neg peak *(better than pos peak if noise and CW signal are combined)**

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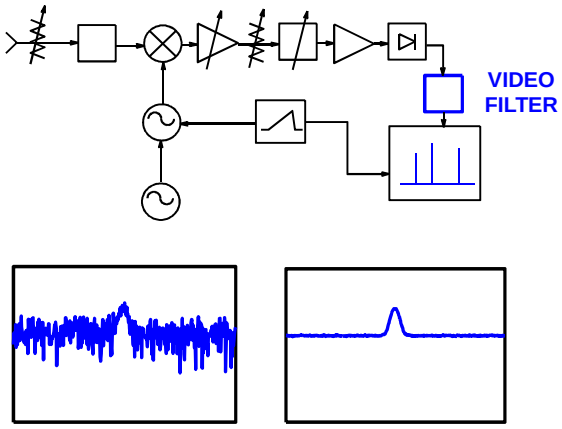
視訊濾波器



Video filter 是低通慮波器，在 detector 之後，在 ADC 之前，可以使顯示的 trace 更平滑。由於曲線上下抖動的幅度變小，原本被淹沒的微弱訊號因而可以較清楚的被呈現。

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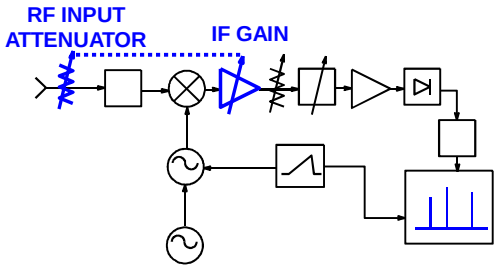
輸入端衰減器與中頻放大器



- 輸入端衰減器為步階式衰減器，低階的 SA 通常每個 step 為 10 dB ，較高階的 SA 可能有5 dB 或 1 dB 的選擇。
- 可防止混頻器過載導致失真，提供混頻器最佳工作點。
- 輸入訊號衰減由 IF section 的放大器自動補償。

本頁取材自 Agilent

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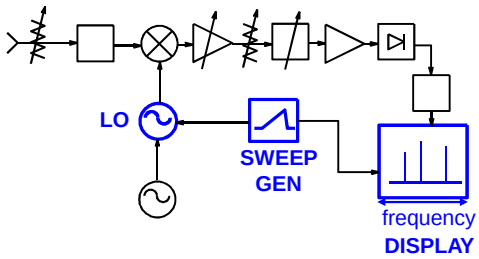
其它(鋸齒波產生器、 本地振盪器)



- 鋸齒波控制顯示螢幕上亮點的水平位置，並根據 SA 面板的頻率設定，控制 local oscillator (LO) 的振盪頻率，校正後，從螢幕上的橫軸即可讀出頻率值。
- LO 可能是 VCO ，也可能是 frequency synthesizer 。
- Frequency synthesizer 的頻率變化是步階式的，其每一個 step 應小於 0.1 RBW ，以防止輸入訊號的某些頻率成份被遺漏或是顯示的 level 出現過大的誤差。

本頁取材自 Agilent

2.4 頻譜分析儀的規格

- Frequency Range
- Accuracy: Frequency and Amplitude
- Resolution (equal and unequal amplitude signals)
- Sensitivity
- Distortion
- Dynamic Range

選擇頻譜分析儀時應該了解

1. 可量測的頻率範圍
2. 被允許的待測訊號強度(最大輸入與靈敏度)
3. 分辨兩待測訊號的強度差異 (Dynamic Range) 與頻率差異 (Resolution)
4. 準確度

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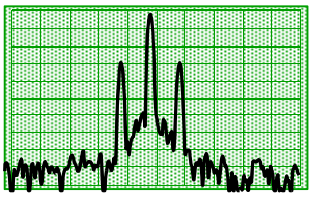
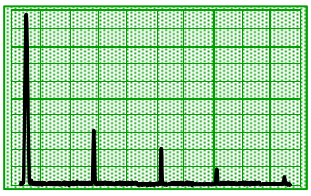
頻率範圍



Low frequencies for baseband and IF



Measuring harmonics 50 GHz and beyond!

許多應用中，除了要知道 fundamental frequency 的訊號強度(或待測元件在 fundamental frequency 的特性)之外，也要知道其諧波成份(或待測元件在 高次諧波的特性) ，尤其是牽涉到無線通訊的電磁干擾問題(要量 10 倍諧波)，或是高速數位系統的訊號完整度時。

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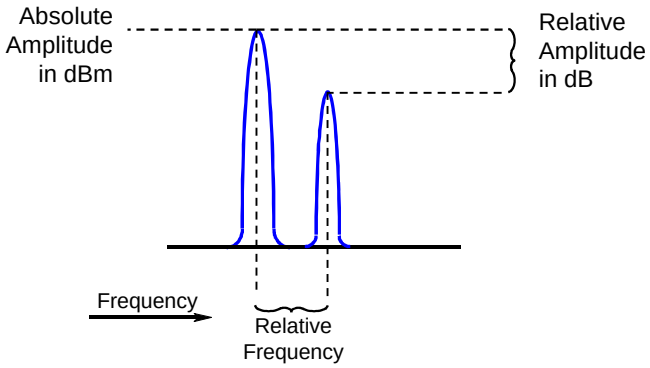
本頁取材自 Agilent

絕對位準與相對位準的量測



- 絕對位準的量測 (如頻率點與 power level) ，可用 single marker 顯示。
- 相對位準的量測 (如相位雜訊、頻道間隔 、頻率漂移) ，可用 delta marker 顯示，通常較絕對位準的量測 還精確。

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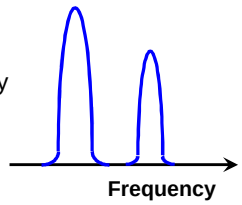
頻率讀值之準確度規格

Typical datasheet specification:

Spans < 2 MHz: ± (freq. readout x freq. ref. accuracy
+ 1% of frequency span
+ 15% of resolution bandwidth
+ 10 Hz "residual error")



- **Frequency Readout 誤差來源包括 Frequency reference inaccuracy 、Span error 、RBW center frequency error 與 FM residual error 。**
- **Frequency reference accuracy 主要由 SA 的架構決定。**
- **Synthesized architecture 之 LO 鎖相至高精度的 溫控 crystal oscillator ，頻率誤差小(幾百 Hz)，價格高。**
- **Free-running architecture之 LO 價格便宜，頻率誤差較大(可能達幾個 MHz) ，價格低。**

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頻率準確度的例子

Single Marker Example: 2 GHz signal
400 kHz span
3 kHz RBW

Calculation:

$(2\times10^{9}$ Hz) x $(1.3\times10^{-7}$ / yr.ref.error)	=	260 Hz
1% of 400 kHz span	=	4000 Hz
15% of 3 kHz RBW	=	450 Hz
10 Hz residual error	=	10 Hz
Total =	±	4720 Hz

從 data sheet 可以找到造成 Frequency ref. inaccuracy 的誤差來源，如
Freq. ref. accuracy = 1 x 10^{-7} (aging) + 0.1 x 10^{-7} (temp stability)
+ 0.1 x 10^{-7} (setability)+ 0.1 x 10^{-7} (warm-up)
= (1.3 x 10^{-7} / yr.) ref. error

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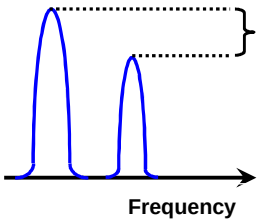
相對訊號強度之準確度



儀器本身特性
- Display fidelity
- Frequency response

可由使用者設定
- Δ RF Input attenuator
- Δ Reference level
- Δ Resolution bandwidth
- Δ Display scaling

- 在做相對性量測時，是以部份 input signal 為 reference 。例如，second harmonic distortion 的量測是以 fundamental harmonic 為基準，只量測 second harmonic 與 fundamental 的強度差異。

相對訊號強度之準確度

- Display fidelity 與 log amplifier的特性、 detector 的線性度以及 digitizing circuit 的線性度有關。
- 在比對兩個訊號相對強度時，最好就兩個訊號個別設定不同的 reference levels ，使兩個訊號在螢幕上都接近 reference level (也就是讓訊號的呈現充滿 screen)，再以 marker 讀出其數值 ， 因為離 reference level 越近的訊號， 其讀值越精確。
- SA 在顯示頻率範圍內的 Frequency response 響應的平坦度越好，相對訊號強度之準確度越佳。
- RF input attenuator 必須工作在較大的頻率範圍， 但 IF gain amplifier 只須工作在 IF 附近， 故通常 RF input attenuator 造成的 inaccuracy 比 IF gain amplifier 大。
- 不同的 RBW 設定，可能使用不同的 IF filter ，其 insertion loss 也不同。
- 改變 display scaling (如 10 dB/div to 5 dB/div) 也會造成誤差。

其它影響訊號強度準確度的因子

- **Mismatch** (RF input port not exactly 50 ohms)
- **Compression due to overload** (high-level input signal)
- **Distortion products**
- **Amplitudes below the log amplifier range**
- **Near noise Signals**
- **Noise causing amplitude variations**
- **Two signals incompletely resolved**

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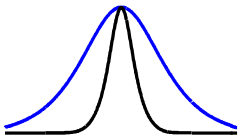
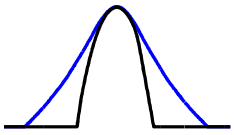
訊號的解析

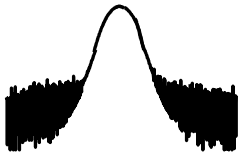


Resolution Bandwidth

Residual FM



IF Filter Type and Shape



Noise Sidebands

- **Resolution (解析度) 指的是分辨兩個頻率很接近的訊號的能力。IF filter 的型態、通帶頻寬與通帶邊緣裙擺的陡峭程度，都會影響解析度。因此 IF filter 的 bandwidth 被稱為 resolution bandwidth (RBW)。**
- **其它影響因素包括 residual FM 與 noise sidebands。**

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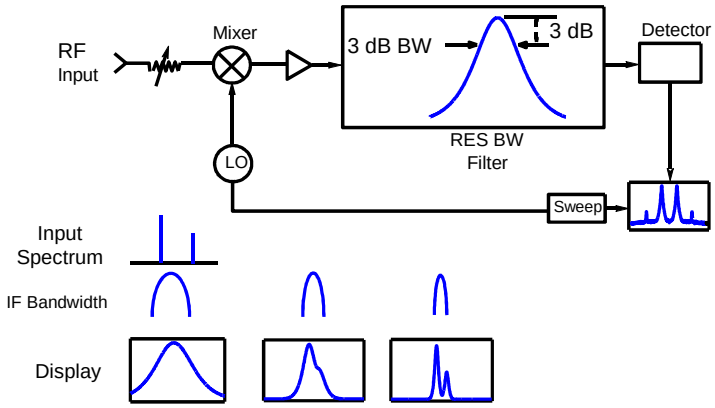
解析頻寬濾波器

IF Filters Allow the Resolution of Adjacent Signals



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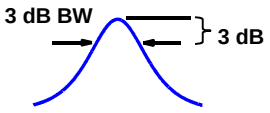
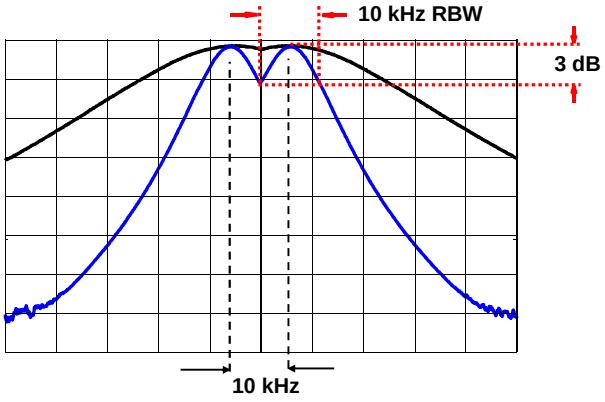
3-dB 解析頻寬—例子

分辨兩個等振幅訊號





兩個等振幅訊號只有在 RBW 小於 兩訊號之頻率間隔時，
才能被辨認。

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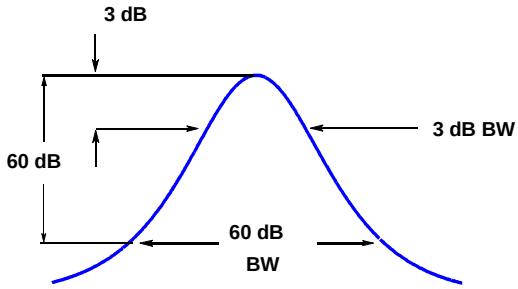
濾波器的選擇性

Determines Resolution of Unequal Amplitude Signals



Selectivity (shape factor) = $\frac{\text{60 dB BW}}{\text{3 dB BW}}$

(Analog filter = 15:1)

(Digital filter = 5:1)

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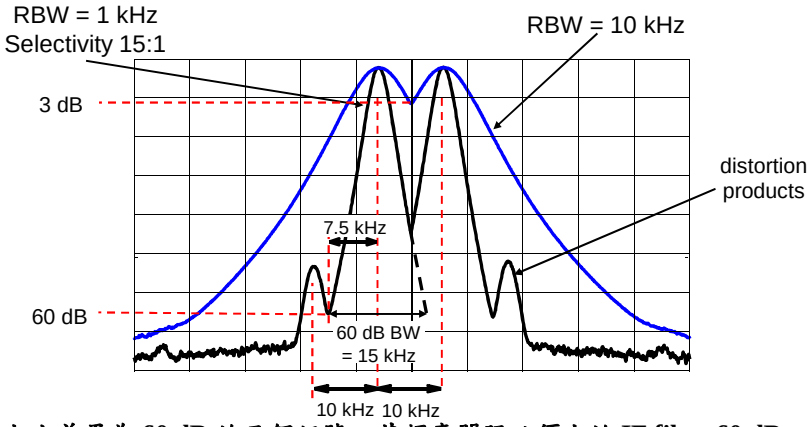
選擇性---例子



大小差異為 60 dB 的兩個訊號，其頻率間隔必須大於 IF filter 60 dB BW 的一半，才能被分辨。以上圖為例，頻率間隔 = 10 kHz ，60 dB BW=15 kHz ，故可被分辨，然而 RBW 才只有 1 KHz 。

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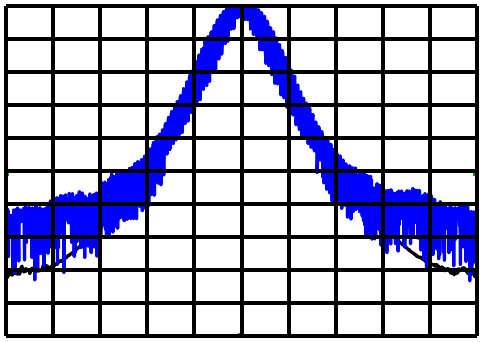
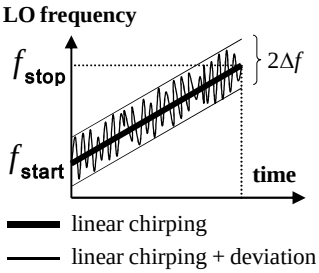
FM 殘留

Residual FM Smears the Signal





若 RBW 小於 LO 的頻率偏移總量(=2 Δf)，則待測的 CW signal 有可能被遺漏，因此 RBW 至少要大於LO 的頻率偏移總量，這個頻率偏移總量也成為兩個CW訊號可以被分辨的最小頻率間隔。使用 frequency synthesizer 的 SA 則無此問題。

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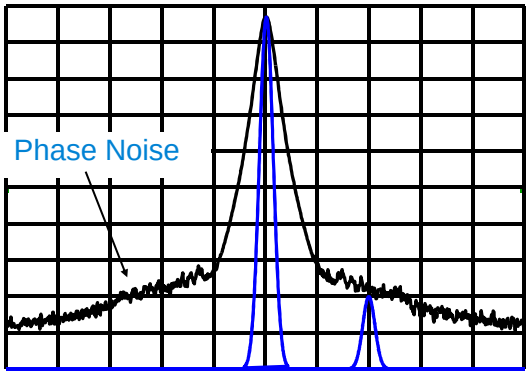
旁波帶雜訊

Noise Sidebands Hide Close-In, Low-Level Signals and prevent resolution of unequal signals



Noise sidebands 源自 LO 的 phase noise

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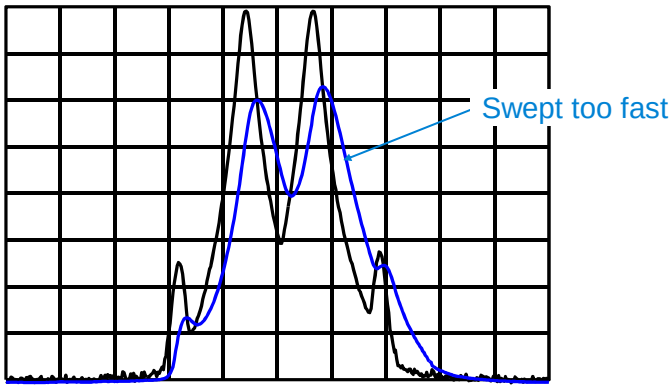
解析頻寬與掃描時間

RBW 是影響掃瞄時間的最重要因子



Penalty For Sweeping Too Fast
Is An Uncalibrated Display

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解析頻寬與掃描時間

- 若 IF filter 的 frequency response 佔據有限的頻率範圍，其對應的 impulse response 佔據無窮長的時間，filter 的輸出是輸入訊號與 impulse response convolve 的結果，因此若要得到完全不失真的輸出，filtering 的動作需要無窮長的時間。
- 實務上 IF filter 是 band-limited 電路，只需幾倍於時間常數的時間進行充放電即可，RBW 越小，所需的充放電時間越長。
- 若掃瞄過快，或掃瞄時間過短，亦即有效的 RBW 大於實際的 RBW ，會造成輸出振幅不足與頻率偏移的失真。
- 若 B = RBW ， R = effective RBW ，T = 掃瞄時間，S = 顯示頻率範圍，且 filer 型態為 Gaussian ，則

$$\frac{R}{B} = \left[1 + 0.195\left(\frac{S}{TB^2}\right)^2\right]^{1/2}$$

$$\text{Amplitude Loss Factor} = \alpha = \left[1 + 0.195\left(\frac{S}{TB^2}\right)^2\right]^{-1/4}$$

本頁取材自 Engelson

解析頻寬與掃描時間

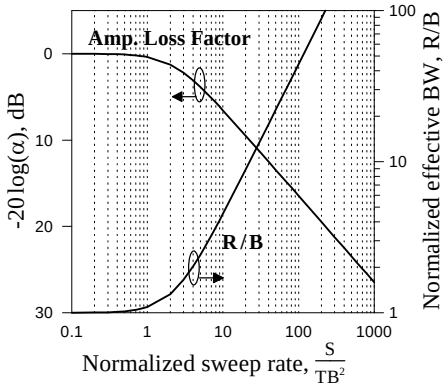
在給定 S 與 T 的前題下，最佳的 RBW 可以求得如下

$$\frac{dR}{dB} = B - \left[\frac{0.195}{B^3}\left(\frac{S}{T}\right)^2\right] = 0$$

$$B_0 = \sqrt{\frac{1}{2.27}\frac{S}{T}}$$

反之，在 S 與 B 設定好之後，最小的掃瞄時間為

$$T = 0.44\frac{S}{B^2}$$



本頁取材自 Engelson

解析頻寬與掃描時間

- 注意，掃瞄時間與 B的平方 成反比，若為了提高解析度而將 RBW 減少為原來的 1/10 ，所需掃瞄時間將提高 100 倍
- 實務上，SA 的掃瞄時間設定更保守

$$T = k\frac{S}{B^2}$$

$$k = \begin{cases} 2 \square 3, & \text{Gaussian filter} \\ 10 \square 15, & \text{Rectangular filter} \end{cases}$$

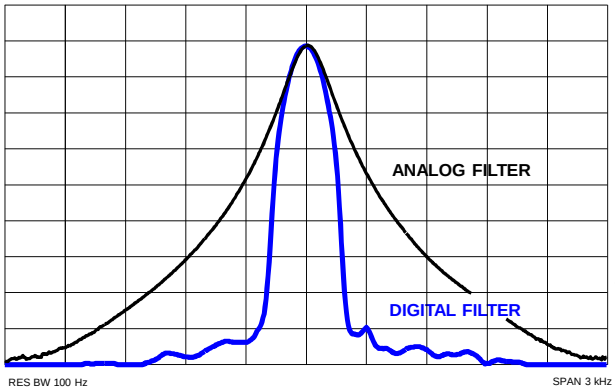
- 若 Video bandwidth = VBW = V 的設定比 RBW 還小，則SA 的掃瞄時間設定將修正為

$$T = k\frac{S}{BV}$$

本頁取材自 Agilent

數位解析頻寬濾波器的優點

Digital RBWs Offer Superior Shape and Measurement Speed



Typical Selectivity

Analog	15:1
Digital	5:1

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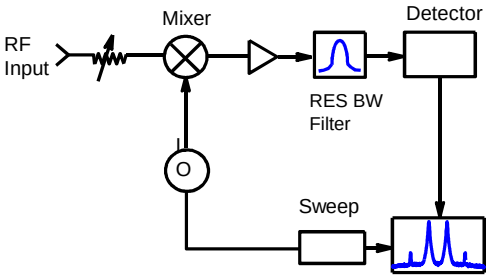
靈敏度與DANL (Displayed Average Noise Level)



- **Spectrum analyzer generates and amplifies noise Just like any other active circuit.**
- **Displayed noise is changed by the RBW.**
- **Displayed average noise level (DANL) is the noise floor of an SA for a given a particular RBW. It represents the best sensitivity of the SA and the ultimate limitation in making measurements on small signals.**

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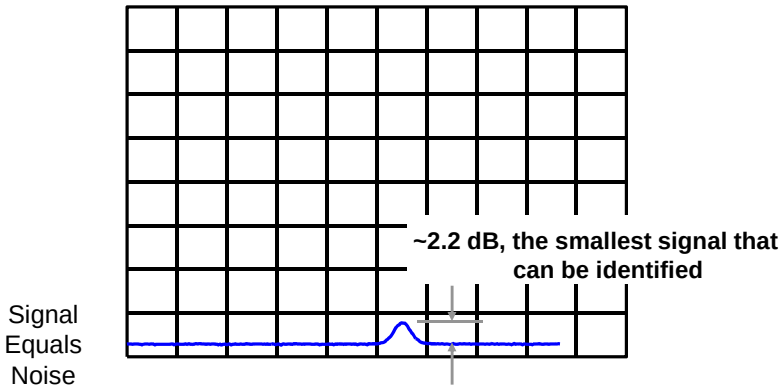
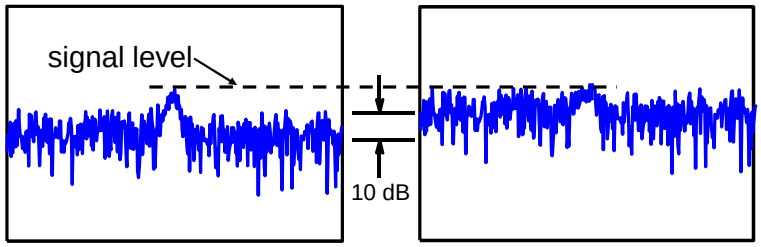
Displayed Average Noise Level

Sensitivity is the Smallest Signal That Can Be Measured

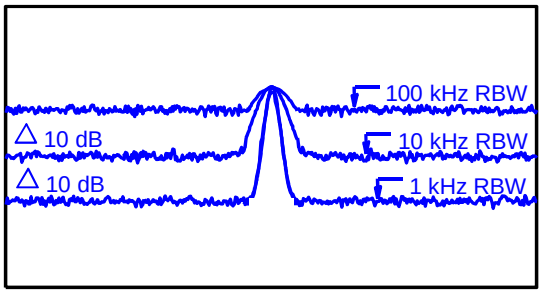


DANL 與 RF 衰減量



- **Effective level of displayed noise is a function of RF input attenuation.**
- **The input signal attenuated by the RF attenuator is compensated by the IF gain amplifier.**
- **Signal-to-noise ratio decreases as RF input attenuation increases.**

DANL 與 RBW



Decreased BW = Decreased Noise

- **Displayed noise is a function of IF filter bandwidth.**
- **If RBW is set to one tenth of its original value, then the DANL is reduced by 10 dB, i.e.,**

  Noise level change (dB) = 10 log (RBW_{new}/RBW_{old})

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最佳靈敏度

- **For best sensitivity, use:**
  - Narrowest RBW
  - Minimum Input Attenuation
  - Appropriate Video Bandwidth Filtering (VBW = 0.01~0.1 RBW) or Video Averaging
- **However**
  - Smaller RBW leads to larger sweep time
  - 0 dB input attenuation causes worse impedance matching at the input

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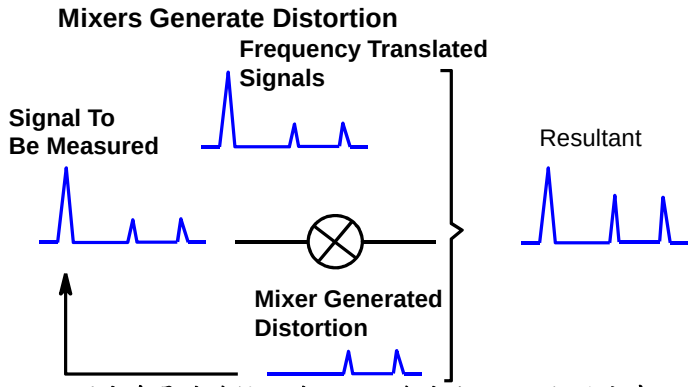
失真



- 混頻器本身是非線性元件，所以會產生 SA 內部的失真，這個內部失真可能在量測待測物的 2 階與 3 階失真時，造成量測誤差。
- 這個內部失真是 Mix Level (即混頻器 RF 端的 power) 的函數，增加 RF input attenuation 可減少此 input power ，並降低內部失真。

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諧波失真 (Harmonic distortion)

一個非線性系統的輸出與輸入之間的關係可表示為

$$v_0(t) = \alpha_1 v_i(t) + \alpha_2 v_i^2(t) + \alpha_3 x_i^3(t) + \cdots$$

若 $v_i(t) = A\cos\omega t$

則

$$v_0(t) = \frac{\alpha_2 A^2}{2} + \left(\alpha_1 A + \frac{3\alpha_3 A^3}{4}\right)\cos(\omega t) + \frac{\alpha_2 A^2}{2}\cos(2\omega t) + \frac{\alpha_3 A^3}{4}\cos(3\omega t)\cdots$$

- 輸入訊號 ~ A → 2階諧波 ~ A^2 → 3階諧波 ~ A^3
- The second harmonic is changed by 2 dB for 1 dB change in input fundamental.
- The third harmonic is changed by 3 dB for 1 dB change in input fundamental.

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互調變失真 (Inter-modulation modulation)

若 $v_i(t) = A\cos\omega_1 t + A\cos\omega_2 t$

則

$$v_0(t) = \left(\alpha_1 + \frac{9}{4}\alpha_3 A^2\right) A\cos\omega_1 t + \left(\alpha_1 + \frac{9}{4}\alpha_3 A^2\right) A\cos\omega_2 t$$
$$+\frac{3}{4}\alpha_3 A^3\cos(2\omega_1 - \omega_2)t + \frac{3}{4}\alpha_3 A^3\cos(2\omega_2 - \omega_1)t + \cdots$$

- **輸入 two-tone 訊號 ~ A**
- **3 階諧波 $(2\omega_1-\omega_2)$ & $(2\omega_2-\omega_1)$ ~ A^3**
- **The third harmonic is changed by 3 dB for 1 dB change in input fundamental.**

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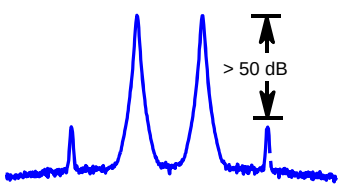
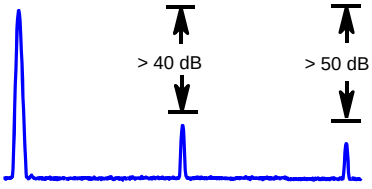
2階與3階失真

Most Influential Distortion is the Second and Third Order



Two-Tone Intermod



Harmonic Distortion

SA 內部產生的 2 階與 3 階失真必須遠小於待測物的失真量時，才能得到可靠的量測結果。

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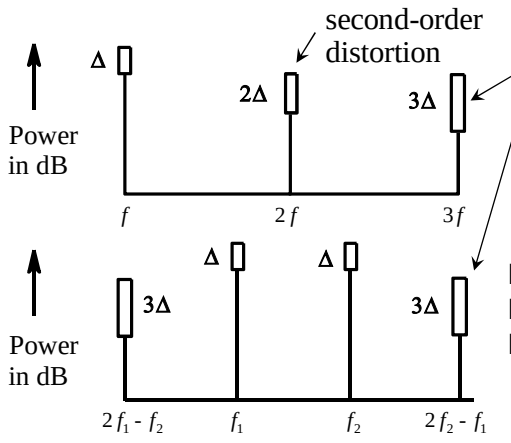
2階與3階失真



Distortion Products Increase as a Function of Fundamental's Power

Second Order: 2 dB/dB of Fundamental
Third Order: 3 dB/dB of Fundamental

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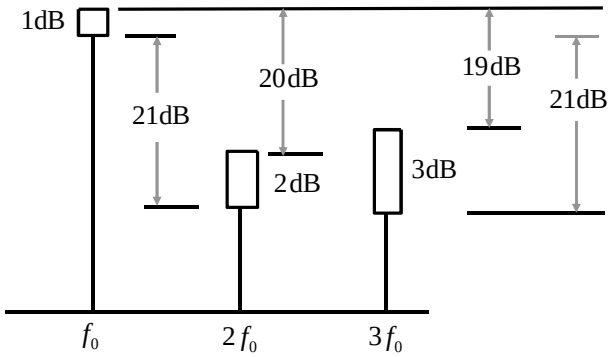
相對訊號強度失真

Relative Amplitude Distortion Changes with Input Power Level



若基頻少 1 dB ，則 2 階失真少 2 dB ， 3 階失真少 3 dB ，但以基頻為基準的相對2 階失真少 1 dB ，相對 3 階失真少 2 dB 。

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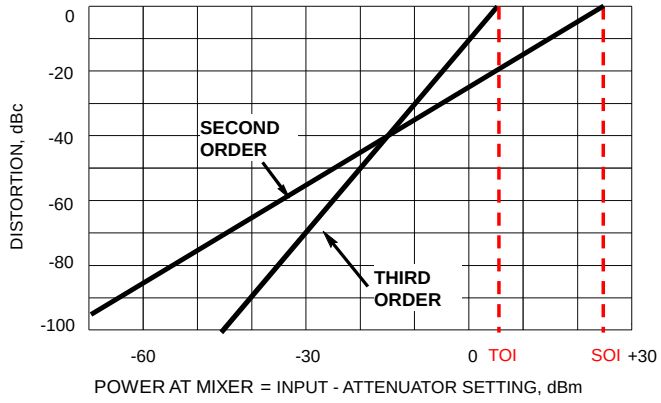
混頻器失真

Distortion is a Function of Mixer Level



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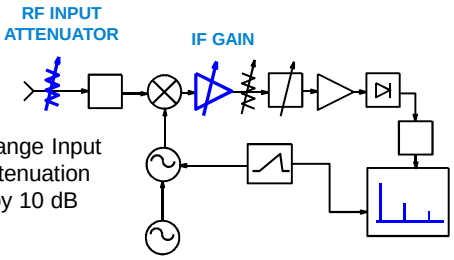
失真測試

Is it Internally or Externally Generated?



1 Change Input Attenuation by 10 dB

2 Watch Signal on Screen:

- No change in amplitude = distortion is part of input signal (external)
- Change in amplitude = at least some of the distortion is being generated inside the analyzer (internal)

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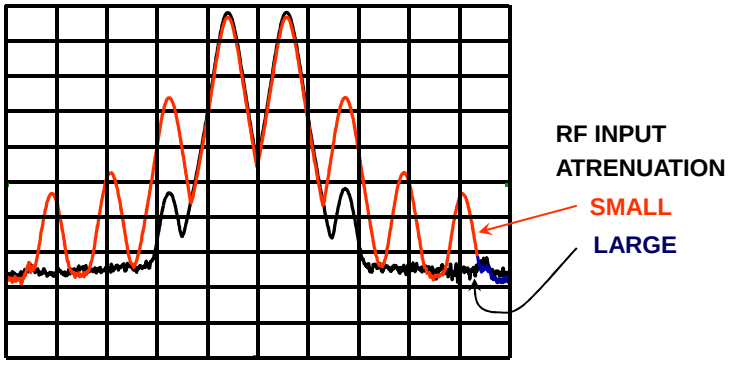
衰減測試



把 RF input attenuation 從大變小，直到螢幕顯示結果即將開始變化為止。

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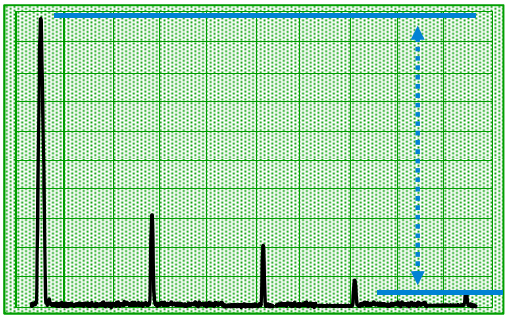
動態範圍



Dynamic Range is the maximum ratio of two signal levels simultaneously present at the input which can be measured to a specified accuracy.

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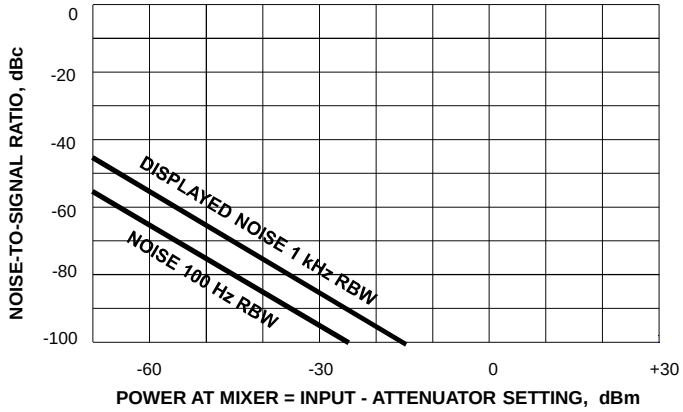
訊噪比

Noise-To-Signal Ratio Can Be Graphed



Best dynamic range occurs at the largest S/N or smallest N/S.

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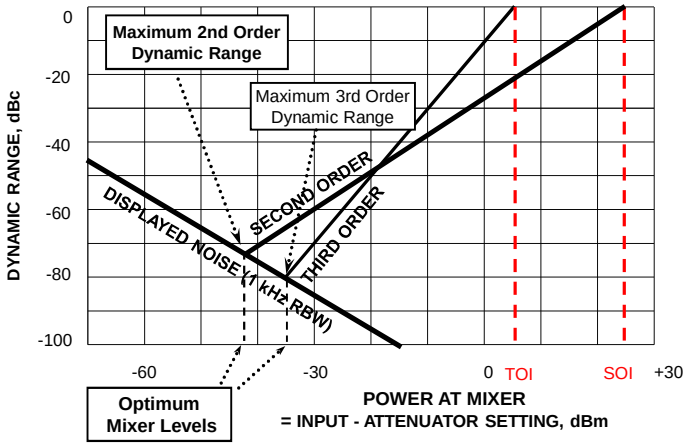
動態範圍圖解

Dynamic Range Can Be Presented Graphically



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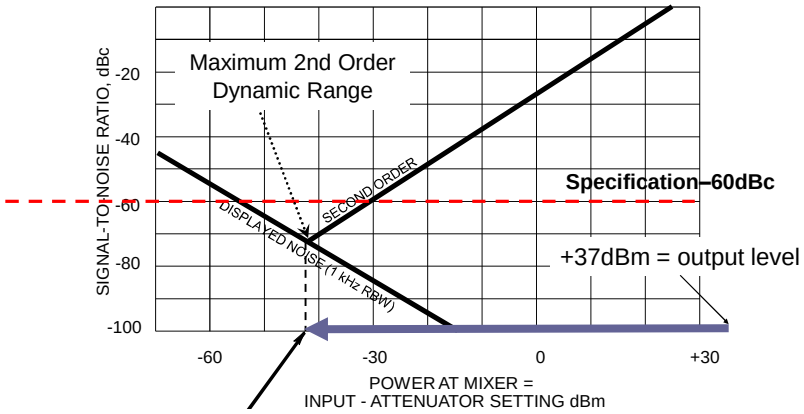
動態範圍– 例子

Use a 40dB external attenuator and 40dB internal,
to get mixer level to –43dBm.



–42dBm = optimum level

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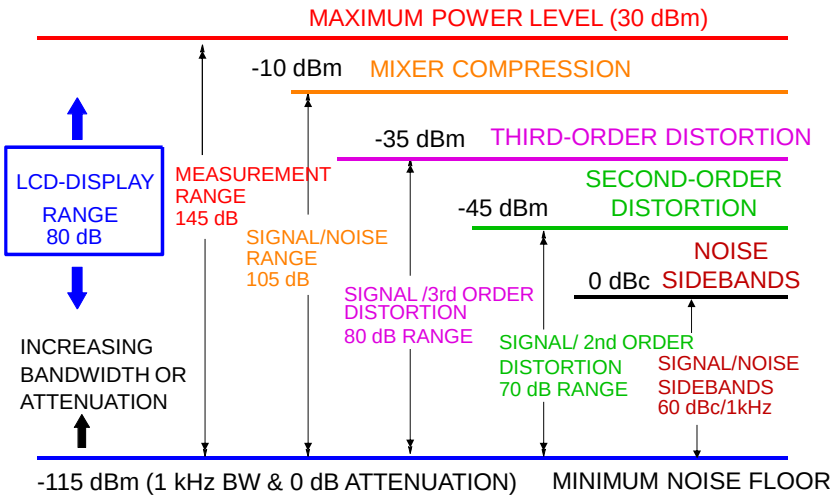
動態範圍– DR Summary Graph



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2.5 量測實例 1 (an LNA circuit)

- **Figure 1 shows a fully integrated 3.1 – 8 GHz UWB LNA.**

Typical schematic of cascode feedback LNA

The modified Chebyshev bandpass filter

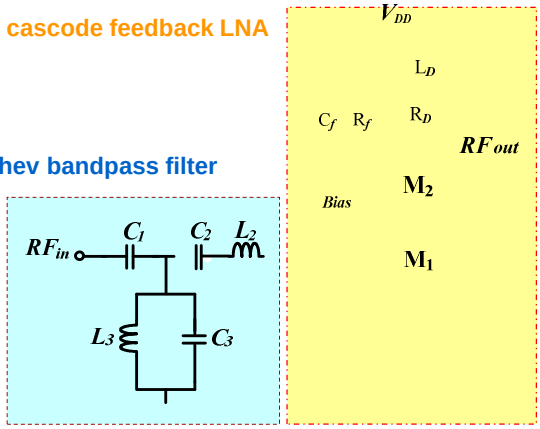


Figure 1 A Fully Integrated 3.1 – 8 GHz Ultra Wideband Low Noise Amplifier

量測實例 1 (an LNA circuit)

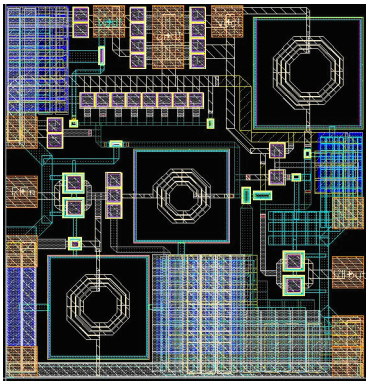


Figure 2 Layout of the proposed UWB LNA

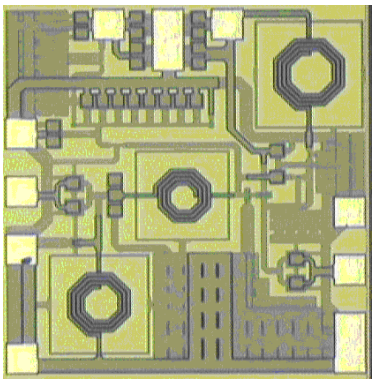
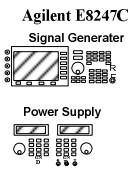
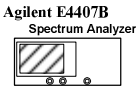


Figure 3 The micrograph without on-wafer measurement

量測實例 1 (an LNA circuit)





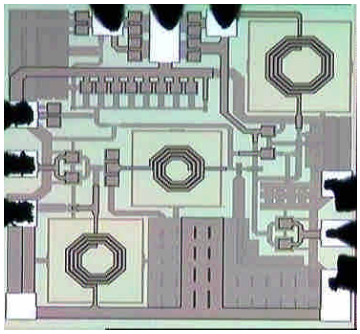


Figure 4 The micrograph with on-wafer measurement

量測實例 1 (an LNA circuit)

- **Figure 5 and 6 show the input P_{1dB} and IIP_3 are -6 dBm and 4 dBm, respectively.**

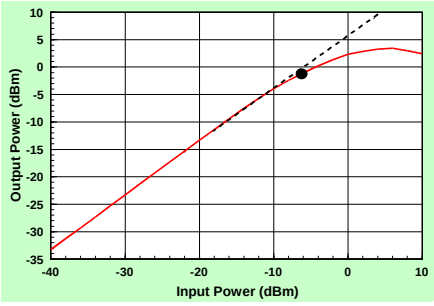


Figure 5 Measured result of P_{1dB}

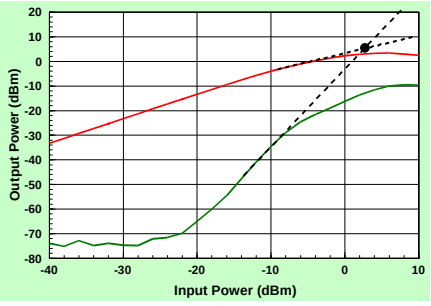


Figure 6 Measured result of IP_3

量測實例 2 (a VCO circuit)

- Figure 7 shows *a fully integrated CMOS 3.5 GHz QVCO*

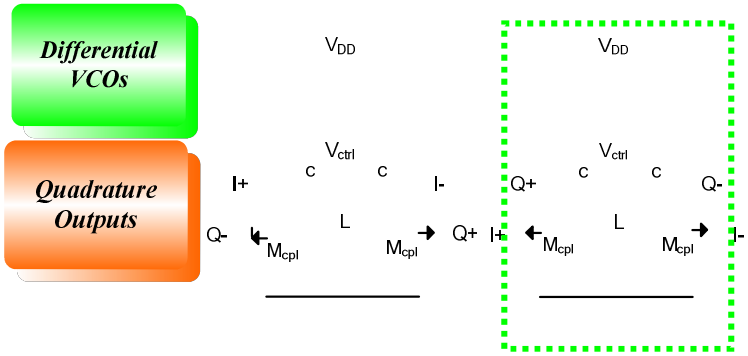


Figure 7 The proposed SQVCO topology

量測實例 2 (a VCO circuit)

- *The layout photograph of the cross-coupled LC QVCO*

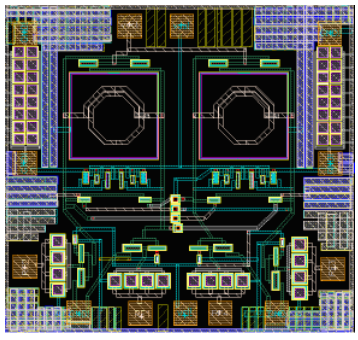


Figure 8 The layout photograph of QVCO

量測實例 2 (a VCO circuit)

- Figure 9 **shows consumption is 15 mA at 2.3 V supplied, and the start-up frequency is 3.776 GHz at .**

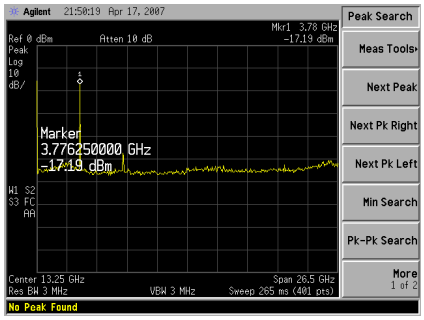


Figure 9 The measurement of the CMOS LC QVCO

量測實例 2 (a VCO circuit)

- **The tuning frequency as shown in *Figure 10* is swept from 0 to 1.8 V.**

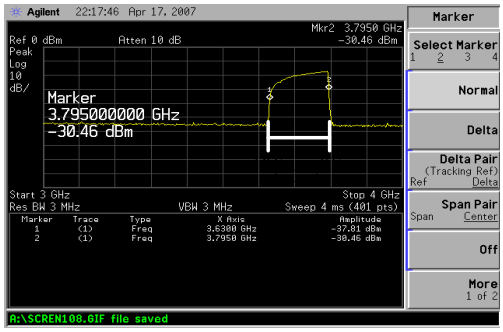
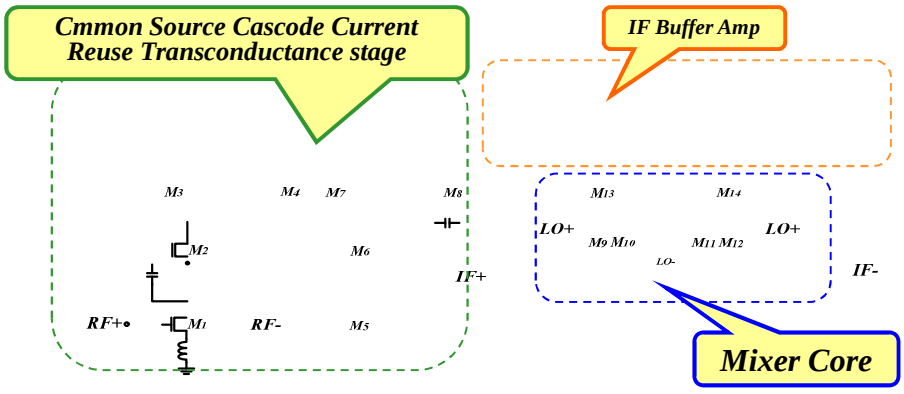


Figure 10 The turning frequency of the CMOS LC QVCO

量測實例 3 (a Mixer circuit)



量測實例 3 (a Mixer circuit)

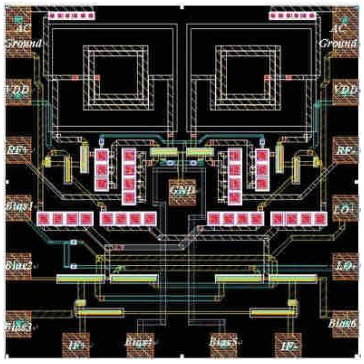


Figure 12 Layout of the proposed Mixer

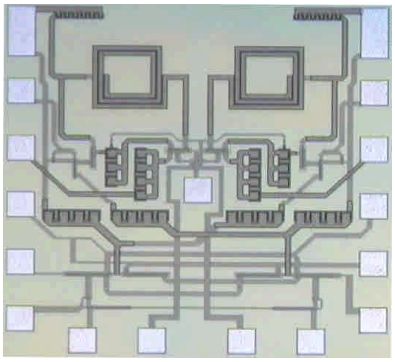


Figure 13 The micrograph without on-wafer measurement

量測實例 3 (a Mixer circuit)

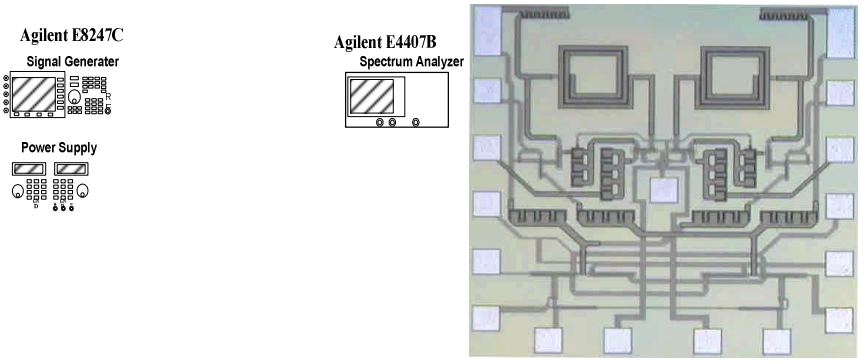


Figure 14 The micrograph with on-wafer measurement

量測實例 3 (a Mixer circuit)

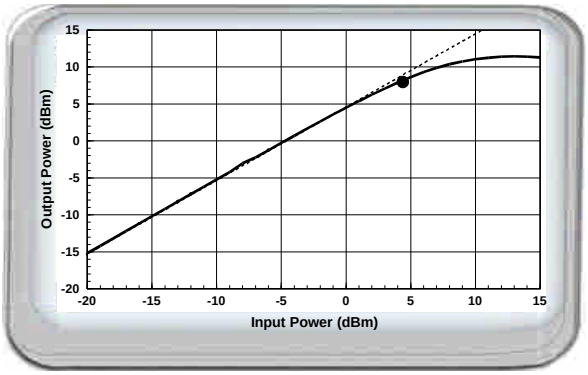


Figure 15 Simulated result of P_{1dB}.

- we can see that the input and output P_{1dB} are about 4.3 dBm and 8.1 dBm.

量測實例 4 (a Wideband Amplifier circuit)

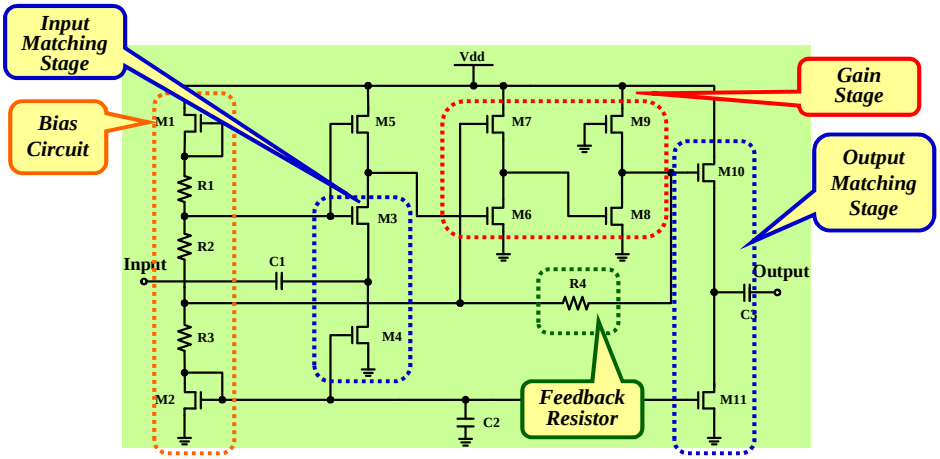


Figure *16 Circuit topology of 0.25 μm wideband amplifier with wide operation voltage range*

量測實例 4 (a Wideband Amplifier circuit)

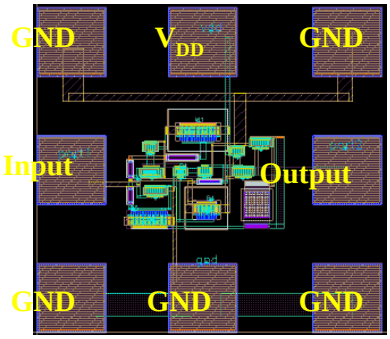


Figure *17 Layout profile of the 0.25 μm wideband amplifier*

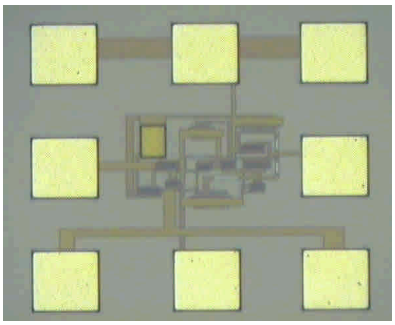


Figure *18 Micrograph of the 0.25 μm wideband amplifier*

量測實例 4 (a Wideband Amplifier circuit)

Measured Results

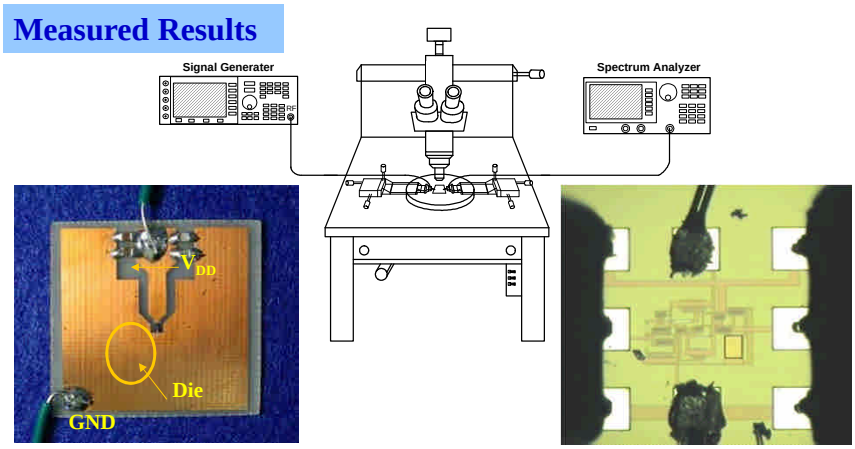


Figure 19 The micrograph with bonding wire PCB measurement

量測實例 4 (a Wideband Amplifier circuit)

2.5 V

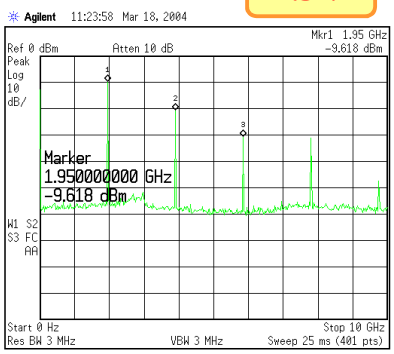


Figure *20 Output spectrum of the 0.25 μm wideband amplifier at 2.5 V supply voltage.*

3 V

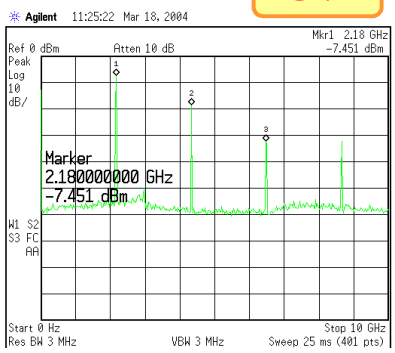


Figure *21 Output spectrum of the 0.25 μm wideband amplifier at 3 V supply voltage.*

量測實例 4 (a Wideband Amplifier circuit)

Figure *22 Output spectrum of the 0.25 μm wideband amplifier at 3.3 V supply voltage.*

Figure *23 Output spectrum of the 0.25 μm wideband amplifier at 4 V supply voltage.*

量測實例 5 (a Power Amplifier circuit)

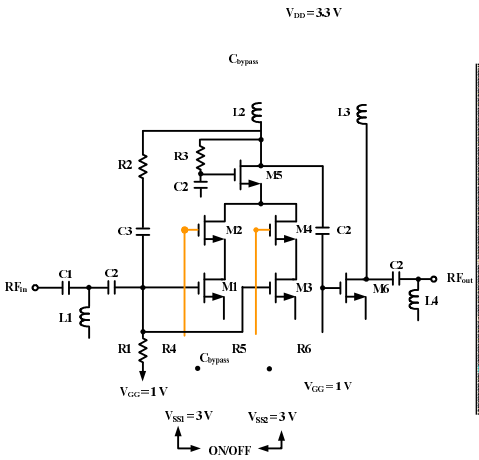


Figure 24 Fully integrated PA

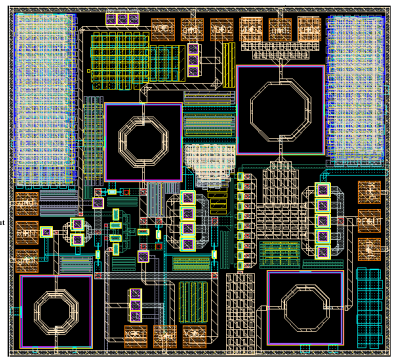


Figure 25 Layout profile of the 0.18 μm power amplifier

量測實例 5 (a Power Amplifier circuit)

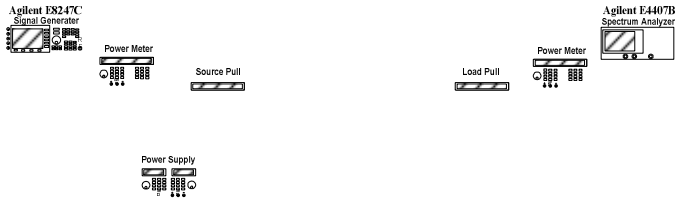


Fig. 26 Measured S_{11}

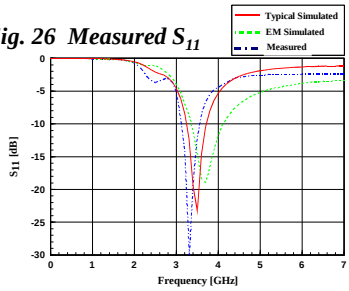
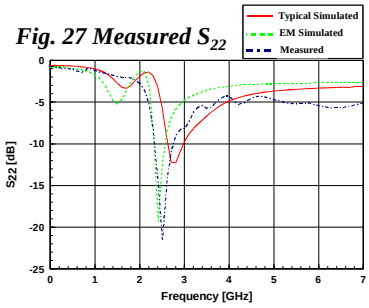


Fig. 27 Measured S_{22}



量測實例 5 (a Power Amplifier circuit)

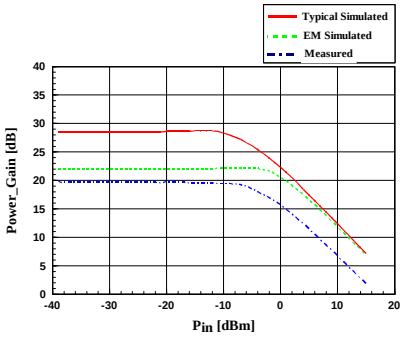


Fig. 28 Measured power gain

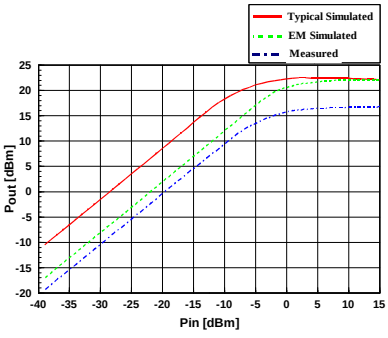
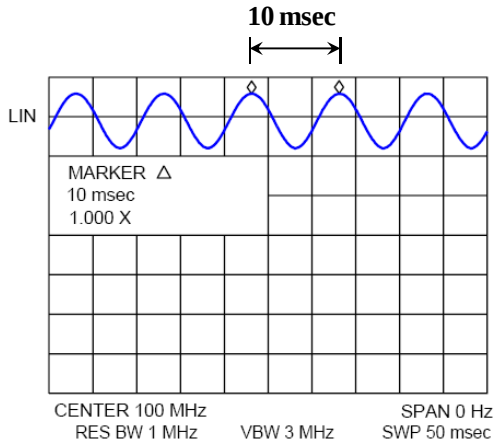


Fig. 29 Measured output power

其它量測配置 1

Zero-Span --- 顯示 modulation signal 的 time-domain 波形



- **Span = 0 Hz**
- **Center frequency = the carrier frequency of the modulated signal**
- **RBW >> f_{mod}，VBW > RBW**
- **Vertical axis = linear**
- **可顯示 modulation signal (baseband signal)**
- **可當 demodulator ，用來 detect AM 訊號 (AM radio)**
- **一般 SA 也配有 FM detector (FM radio)**
- **左圖例子，調變訊號頻率 = f_{mod} = 1/T = 100 Hz**

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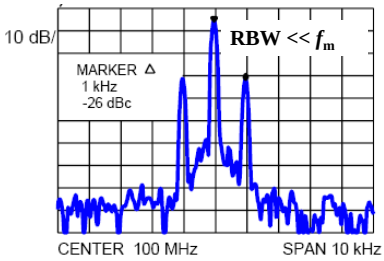
其它量測配置 1

Zero-Span --- 可決定 AM signal 的 modulation index

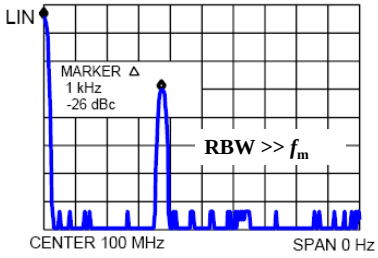
Swept Frequency Domain



$$v_{AM}(t) = (1 + m\cos\omega_m t)\cos\omega_c t$$

modulation index = $2\times10^{A_{dB}/2}$

Zero Span + FFT Frequency Domain



sideband level relative to carrier
= $A_{dB} = 20\log_{10}(m/2)$

For this example, m = 0.1 or 10%

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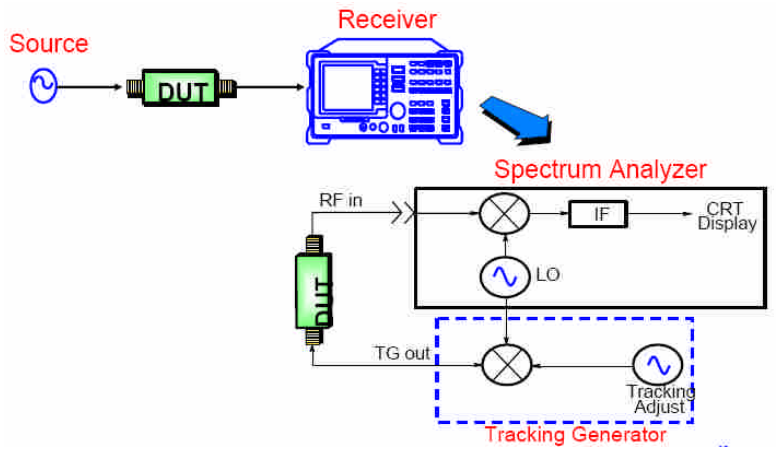
其它量測配置 2



上圖配置可量 S_{21} ，若搭配外部 directional coupler 也可量 S_{11}

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