Sub-6GHz/mmW共構之多輸入輸出射頻模組－射頻傳收機架構

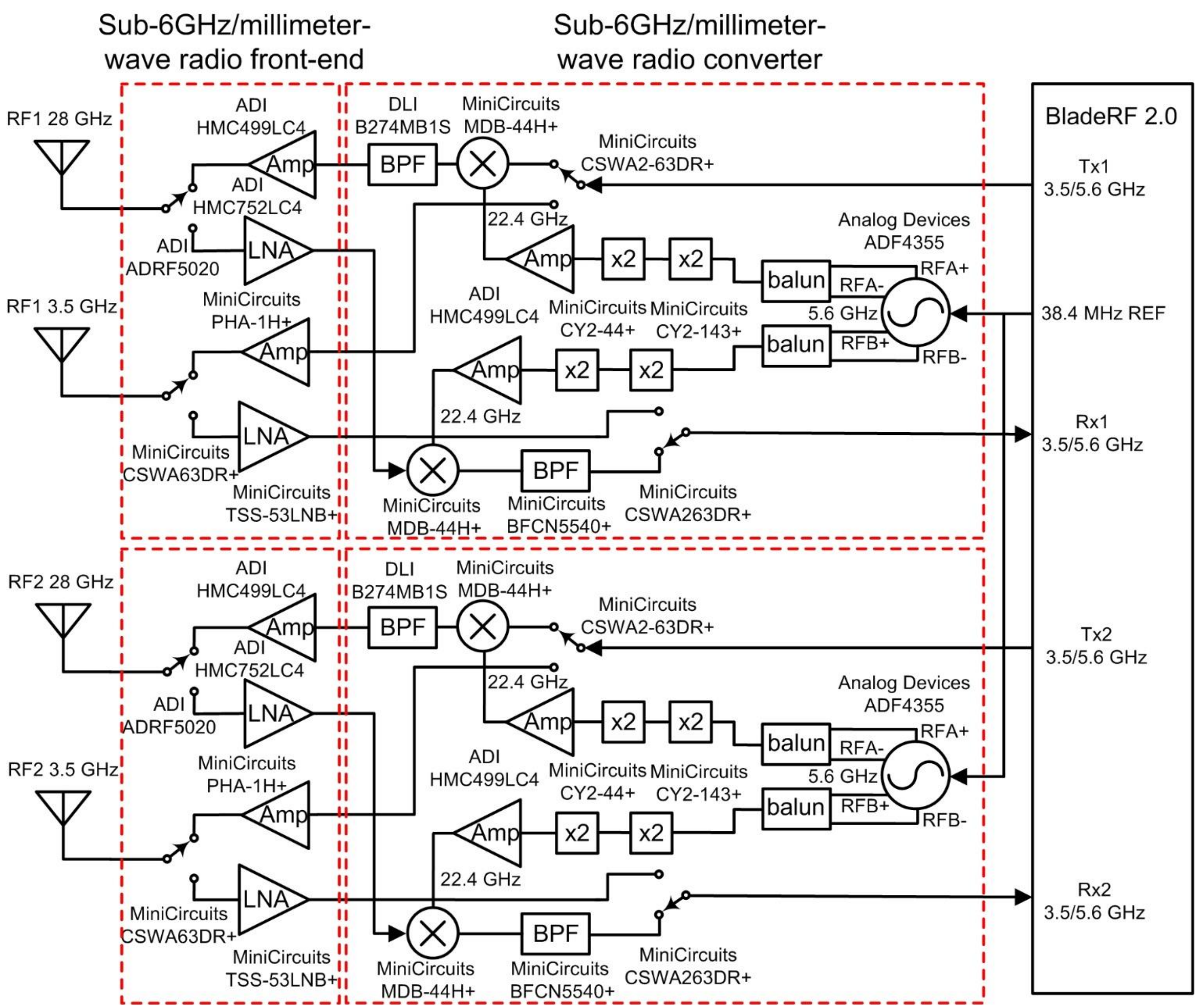
大綱

- Sub-6GHz/mmW共構之多輸入輸出射頻傳收機
  - 設計考量
  - Sub-6GHz軟體無線電平台
  - Sub-6GHz/mmW轉換器模組
  - Sub-6GHz/mmW射頻前端模組
  - 天線模組
- 多輸入輸出射頻傳收機組成元件
  - 頻率合成器
  - 頻率乘法器
  - 升／降頻混頻器
  - 濾波器
  - 低雜音放大器
  - 功率放大器
  - 射頻開關
  - 天線

Sub-6GHz/mmW射頻傳收機架構設計考量

- 需展示sub-6GHz/mmW多輸入輸出功能
- 盡量利用具多輸入輸出之商用軟體無線電平台
- 軟體無線電平台需支援常用之軟體系統
- 具擴充性
- 使用低成本且已封裝完成之微波／毫米波元件
- 除非測試需求，不使用價格高昂的K-type接頭
- 天線採印刷電路板製作之平面式天線
- 印刷電路板利用半導體研究中心提供之製程服務

Sub-6GHz/mmW 2x2射頻傳收機架構圖



Sub-6GHz/mmW 2x2射頻傳收機架構說明

- 射頻傳收架構概分為軟體無線電、射頻轉換器與射頻前端三大區塊
- 利用具2收2發之商用BladeRF-2.0軟體無線電平台
- 以MATLAB做軟體發展平台與介面
- 可展示3.5 GHz或28 GHz之射頻傳收
- 可展示SISO或2x2-MIMO功能
- 所有IC皆為封裝完成之商用產品
- 28 GHz採用平面式八木天線， 3.5 GHz採用微帶天線
- 印刷電路板為半導體研究中心提供之RO4003C/FR4四層板製程服務

Sub-6GHz BladeRF 2.0軟體無線電平台

- ADC/DAC sample rate: 0.521 MSPS – 61.44 MSPS
- ADC/DAC resolution: 12 bits
- VCTCXO calibrated accuracy: 16 ppb
- RF Rx tuning range: 70 MHz – 6 GHz
- RF Rx tuning range: 47 MHz – 6 GHz
- RF bandwidth filter: 0.2 MHz – 56 MHz
- CW output power: 8 dBm
- FPGA logic elements: 49 KLE – 301 KLE
- FPGA memory: 3,383 Kbits – 13,917 Kbits
- Variable-precision DSP blocks: 66 – 342
- Embedded 18x18 multiplier: 132 – 684

Source: https://www.nuand.com/bladerf-2-0-micro/

Sub-6GHz/mmW轉換器模組

- 頻率合成器
- 頻率乘法器
- 本地振盪信號放大器
- 升／降頻混頻器
- 帶通濾波器

Sub-6GHz/mmW射頻前端模組

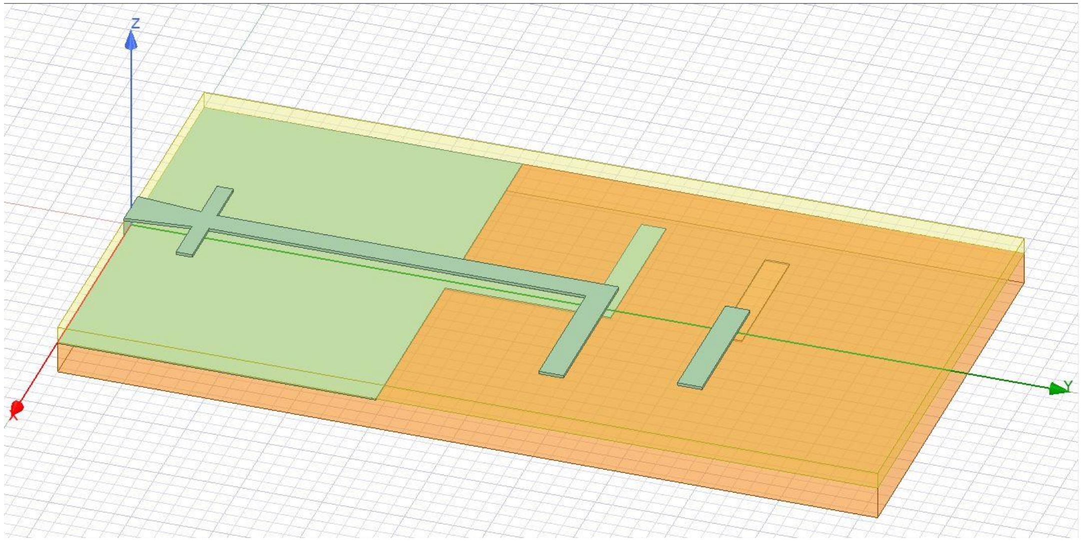
- 低雜音放大器
- 功率放大器
- 射頻開關

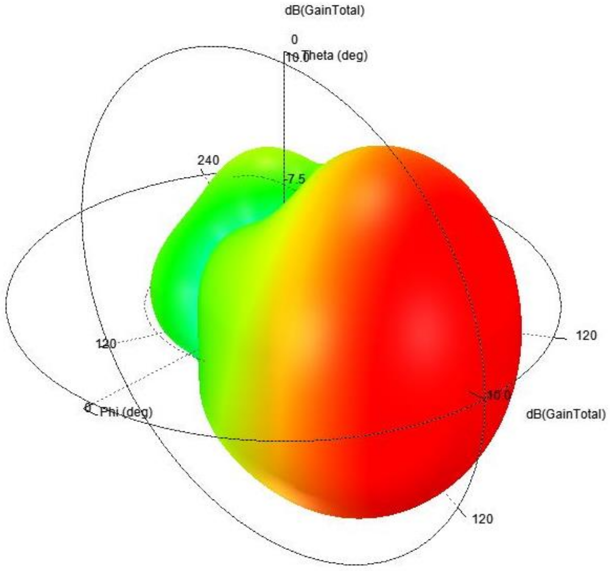
天線模組

- 3.5 GHz 1x2 微帶天線陣列
- 28 GHz 1x2 平面式八木天線陣列

天線模組 - I

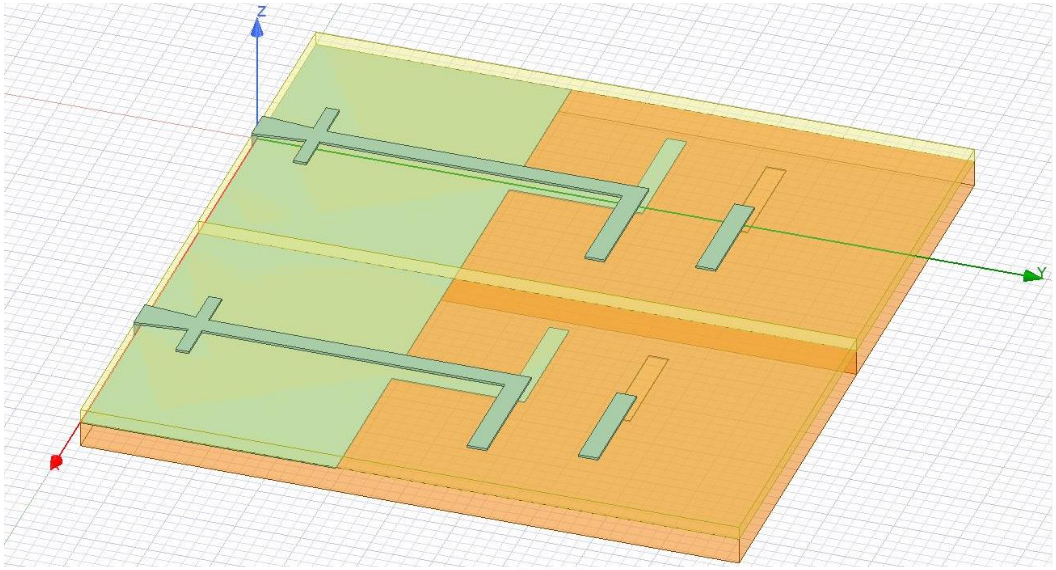
- RO4003C/FR4四層板
- 平面式八木天線
- End-fire輻射型式
- 增益7.4 dB
- 反射係數 < -10dB頻寬 1 GHz以上
- 面積5.36mm x 10mm

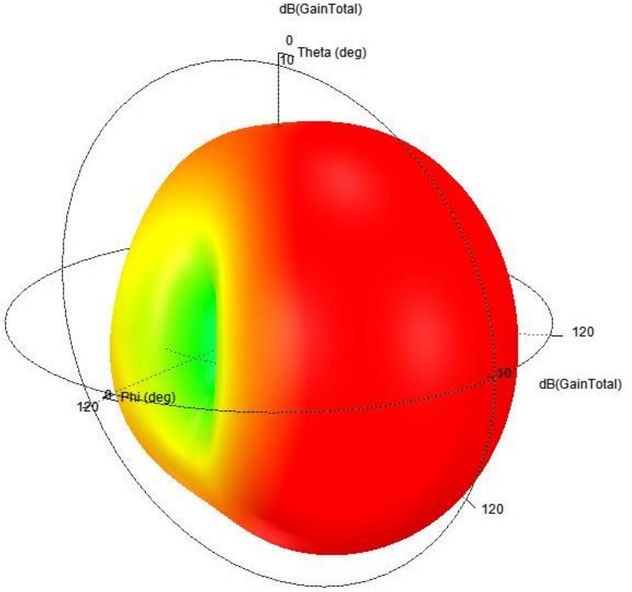




天線模組 - II

- RO4003C/FR4四層板
- 平面式八木天線
- End-fire輻射型式
- 增益8.0 dB
- 反射係數 < -10dB頻寬 1 GHz以上
- 面積10.72mm x 10mm





Sub-6GHz/mmW共構之多輸入輸出射頻模組－射頻傳收機電路模組分析、設計與量測

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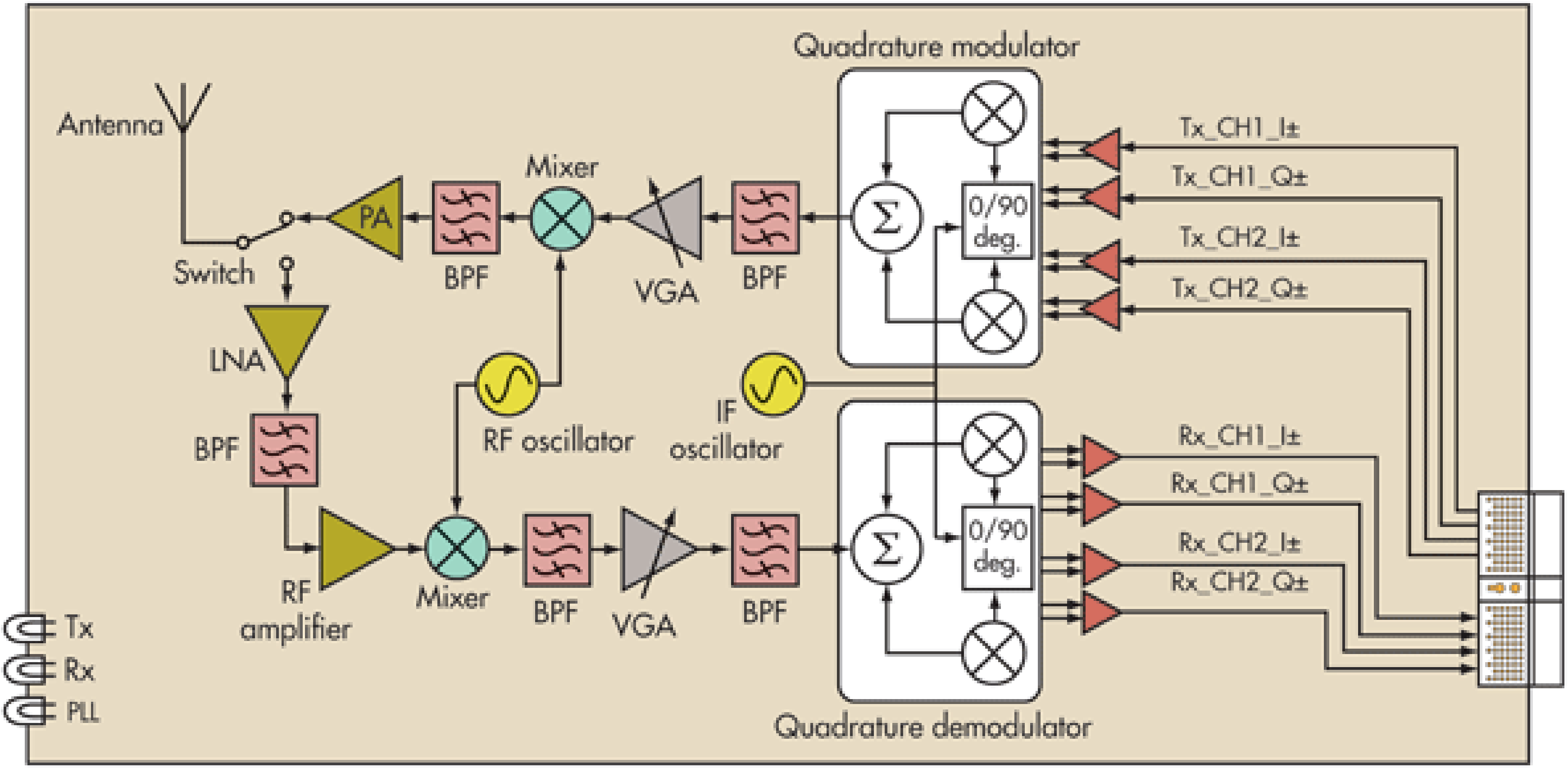
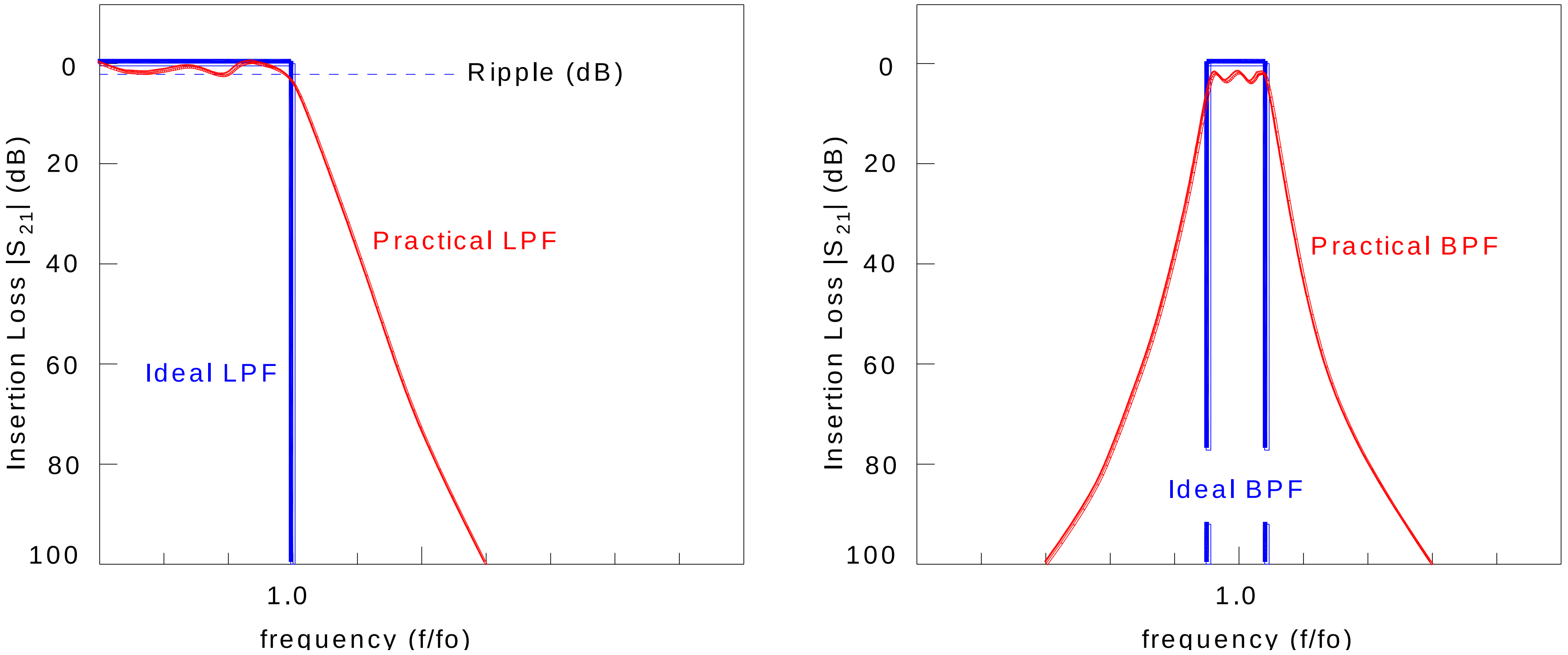
The role of RF filters



Microwave filters

- Filter types: Low-pass, High-pass, Band-pass, Band-reject, All-pass, ...



Design methods

(1) Image Parameter Method

- A cascade of simple two-port filter sections to provide the desired cutoff
- frequencies and attenuation characteristics, but not allow the spec of a frequency response over the complete operating range.

(2) Insertion Loss Method

- Network synthesis technique to complete the whole frequency response.

Design Procedure

Filter Specification	N (order), f_o, Δf, or $\Delta f/f_o$ (fractional bandwidth), Ripple (dB), minimal attenuation at f_1/f_o, …, etc.
⇩	
Low-pass Prototype	A prescribed filter function with Z_o normalized to 1 Ω, and ω normalized to 1 rad./sec. Order N, Δ, and ripple level are required.
⇩	
Scaling Conversion	Frequency and impedance scaling are required for transforming the prototype function to the desired filter function.
⇩	
Implementation	Filters of Microstrip, CPW, Lumped elements and Rectangular waveguide are implemented in different ways. Their theories are the same.

Coupled line filters

Design formulas for a coupled-line filter

Once J_1, J_2, and J_3 are found, then Z_{0e} and Z_{0o} can be determined via

$$\begin{cases} Z_{0e} = Z_0\left[1 + JZ_0 + (JZ_0)^2\right] \\ Z_{0o} = Z_0\left[1 - JZ_0 + (JZ_0)^2\right] \end{cases}$$

For any N and arbitrary Z_L (Z_L = Z_0 (g_{N+1} = 1) or $Z_L \neq Z_0$ ($g_{N+1} \neq 1$)),

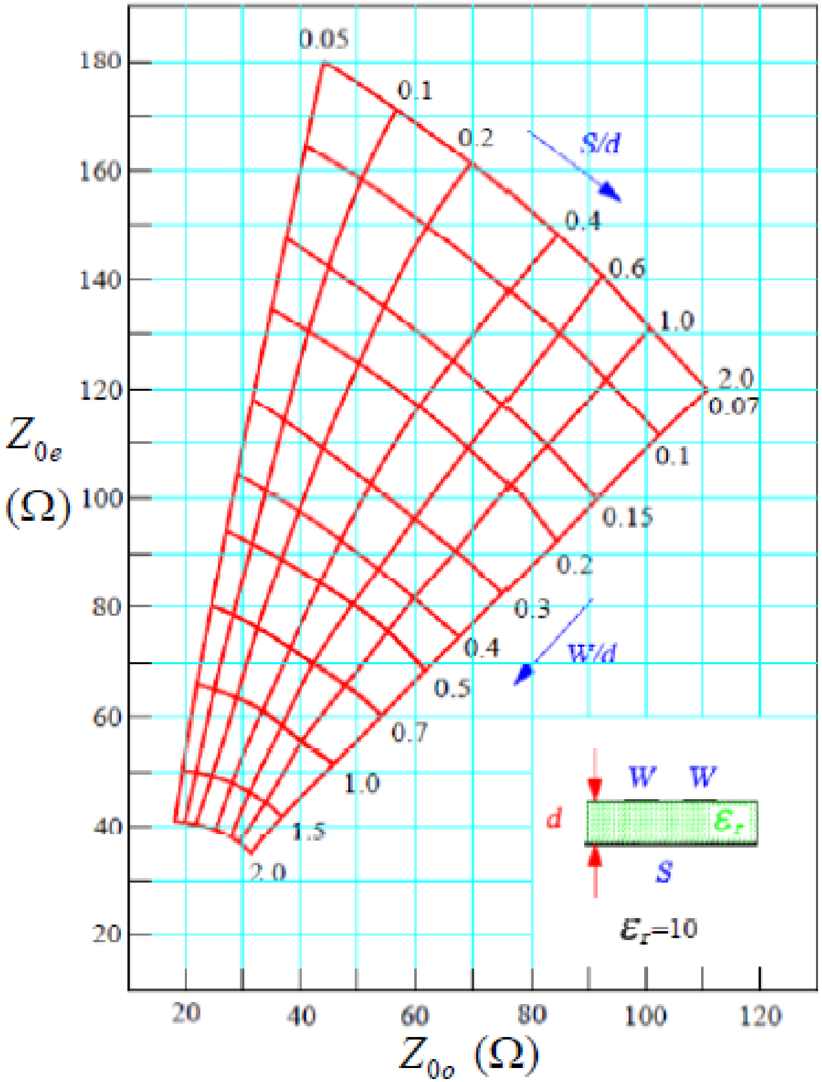
Design data:
StripLine: w_n/d and s_n/d
Microstrip: w_n/d and s_n/d

$$Z_0 J_1 = \sqrt{\frac{\pi\Delta}{2g_1}}$$

$$Z_0 J_n = \frac{\pi\Delta}{2\sqrt{g_{n-1}g_n}}, \quad n = 2, 3, \cdots, N$$

$$Z_0 J_{N+1} = \sqrt{\frac{\pi\Delta}{2g_N g_{N+1}}}$$

Z_{oe} and Z_{oo} Design Data for a Microstrip



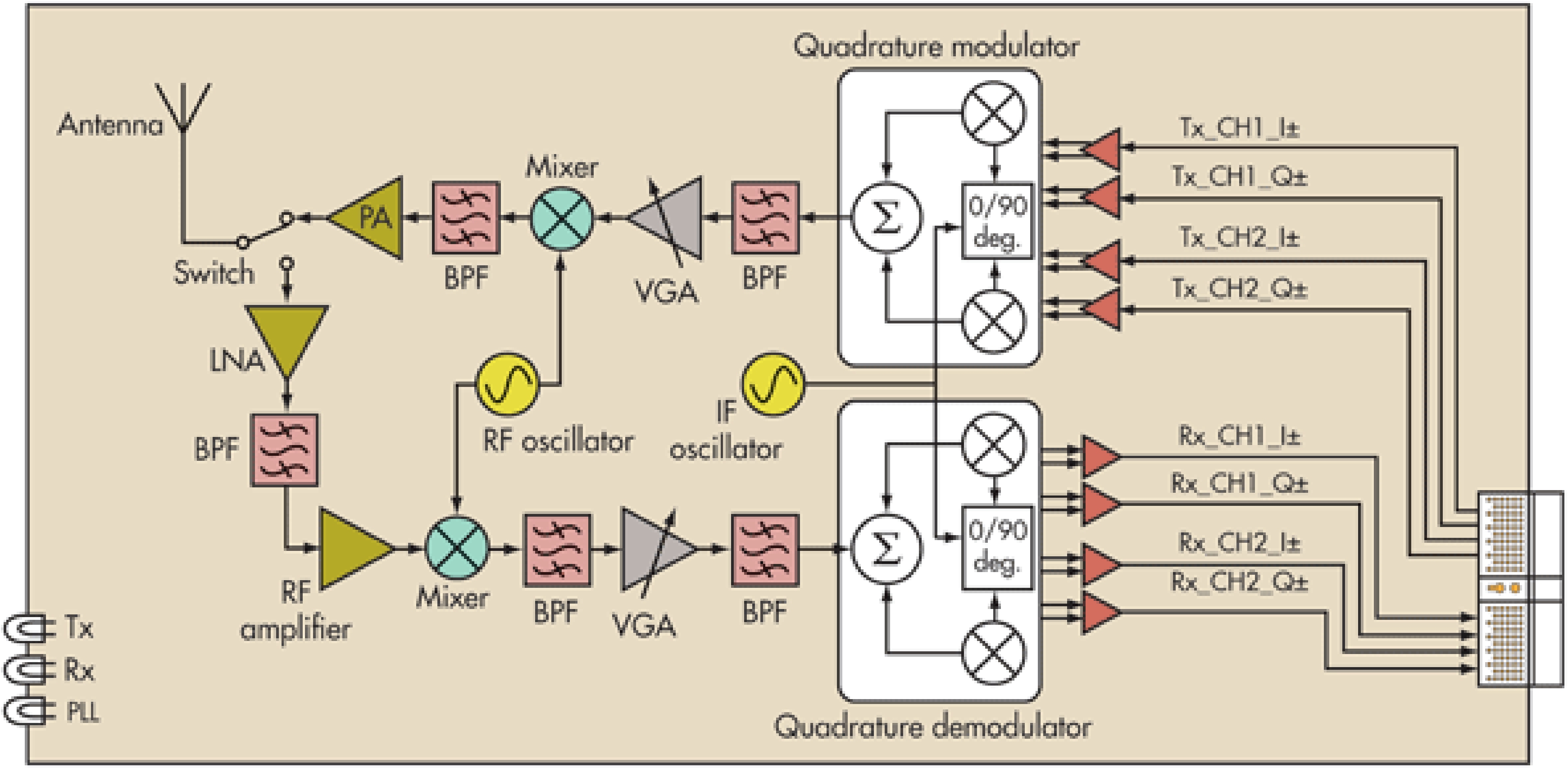
(1) Quasi-static results, realistic microstrip is dispersive. Usable up to 5–6 GHz.

(2) If ε_r is different or frequency is sufficiently high, the data must be changed.

(3) Given Z_{0e} and Z_{0o}, s/d and w/d are uniquely specified.

(4) Phase velocity data is not shown.

The role of RF amplifiers



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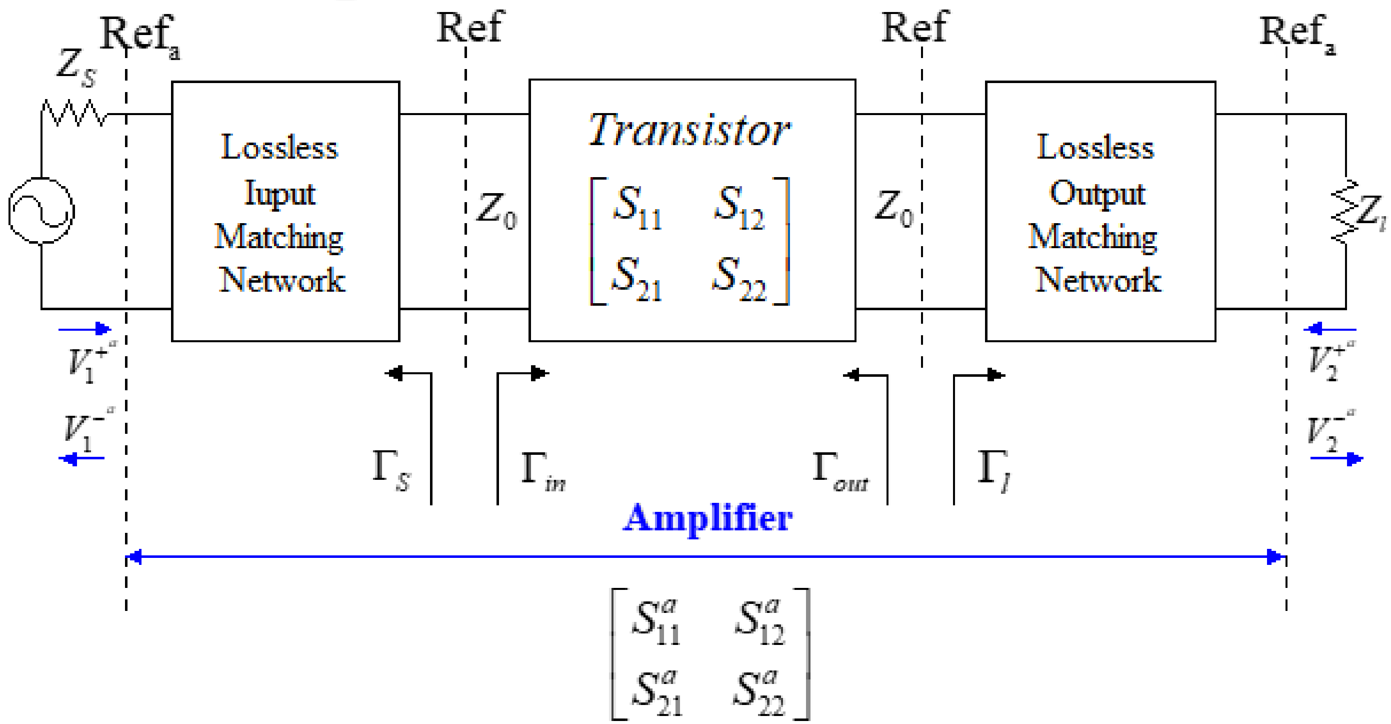
Microwave Amplifier Gain



$$P_{avs} = \frac{\left|V_1^{+a}\right|^2}{2Z_0}, \quad P_l = \frac{\left|V_2^{-a}\right|^2}{2Z_0} \Rightarrow \left|S_{21}^a\right|^2 = \frac{\left|V_2^{-a}\right|^2}{\left|V_1^{+a}\right|^2}\Bigg|_{V_2^{+a}=0} = \frac{P_l}{P_{avs}} = G_T = \frac{\left|S_{21}\right|^2\left(1-\left|\Gamma_S\right|^2\right)\left(1-\left|\Gamma_l\right|^2\right)}{\left|1-S_{22}\Gamma_l\right|^2\left|1-\Gamma_S\Gamma_{in}\right|^2}$$

- For lossless input and output matching networks, the amplifier gain $|S_{21}^a|^2$ is equal to the transistor transducer gain (G_T).

Other Power Gain

$$G_T(\Gamma_S,\Gamma_l)=\frac{|S_{21}|^2(1-|\Gamma_S|^2)(1-|\Gamma_l|^2)}{|1-S_{22}\Gamma_l|^2|1-\Gamma_S\Gamma_{in}|^2}=\frac{(1-|\Gamma_S|^2)|S_{21}|^2(1-|\Gamma_l|^2)}{|(1-S_{11}\Gamma_S)(1-S_{22}\Gamma_l)-S_{12}S_{21}\Gamma_S\Gamma_l|^2}$$

$G_{TU}(\Gamma_S,\Gamma_l)=$ Unilateral transducer power gain

$$=G_T(\Gamma_S,\Gamma_l)\Big|_{S_{12}=0}=\frac{(1-|\Gamma_S|^2)}{|1-S_{11}\Gamma_S|^2}|S_{21}|^2\frac{(1-|\Gamma_l|^2)}{|1-S_{22}\Gamma_S|^2}$$

$G_P(\Gamma_l)=$ Operating power gain

$$=G_T(\Gamma_S=\Gamma_{in}^*,\Gamma_l)=\frac{|S_{21}|^2(1-|\Gamma_l|^2)}{(1-|\Gamma_{in}|^2)|1-S_{22}\Gamma_l|^2}$$

$G_A(\Gamma_S)=$ Available power gain

$$=G_T(\Gamma_S,\Gamma_l=\Gamma_{out}^*)=\frac{|S_{21}|^2(1-|\Gamma_S|^2)}{(1-|\Gamma_{out}|^2)|1-S_{11}\Gamma_S|^2}$$

$$G_{max} = \text{Maximum available power gain}$$

$$= G_T\left(\Gamma_S = \Gamma_{in}^*, \Gamma_l = \Gamma_{out}^*\right) = \frac{|S_{21}|}{|S_{12}|}\left(k - \sqrt{k^2 - 1}\right), \quad k \text{ is stability factor}$$

$$G_{TU\,max} = \text{Maximum unilateral power gain}$$

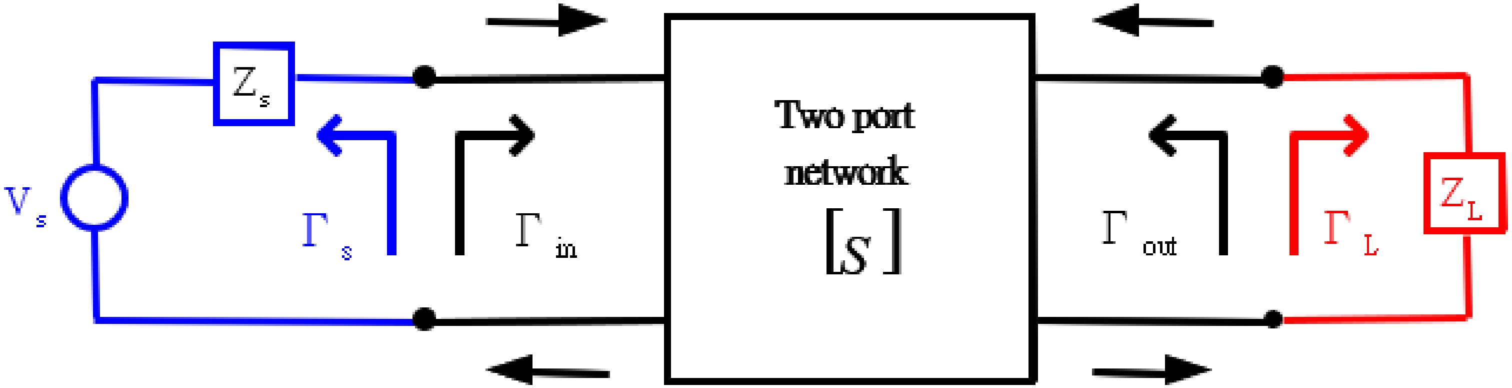
$$= G_{TU}\left(\Gamma_S = \Gamma_{in}^*, \Gamma_l = \Gamma_{out}^*\right) = \frac{|S_{21}|^2}{\left(1 - |S_{11}|^2\right)\left(1 - |S_{22}|^2\right)}$$

$$G_{ms} = \text{Maximum stable power gain}$$

$$= G_{max}\big|_{k=1} = \frac{|S_{21}|}{|S_{12}|}$$

- In general, the amplifier gain will be larger than the transistor gain due to the impedance matching of the transistor at both input and output.

Stability Condition



Unconditional stable:

$$\forall \Gamma_S, \Gamma_L \rightarrow$$

$$|\Gamma_{in}| = \left| S_{11} + \frac{S_{21}\Gamma_L S_{12}}{1 - S_{22}\Gamma_L} \right| < 1$$

$$|\Gamma_{out}| = \left| S_{22} + \frac{S_{21}\Gamma_L S_{12}}{1 - S_{11}\Gamma_S} \right| < 1$$

Conditional stable:

$$\exists \Gamma_S, \Gamma_L \rightarrow$$

$$|\Gamma_{in}| = \left| S_{11} + \frac{S_{21}\Gamma_L S_{12}}{1 - S_{22}\Gamma_L} \right| > 1$$

$$|\Gamma_{out}| = \left| S_{22} + \frac{S_{21}\Gamma_L S_{12}}{1 - S_{11}\Gamma_S} \right| > 1$$

Determine Unconditional stability by mathematical equations

- Rollet's condition (k-Δ test) for unconditional stability.

$$k = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2|S_{12}S_{21}|} > 1$$

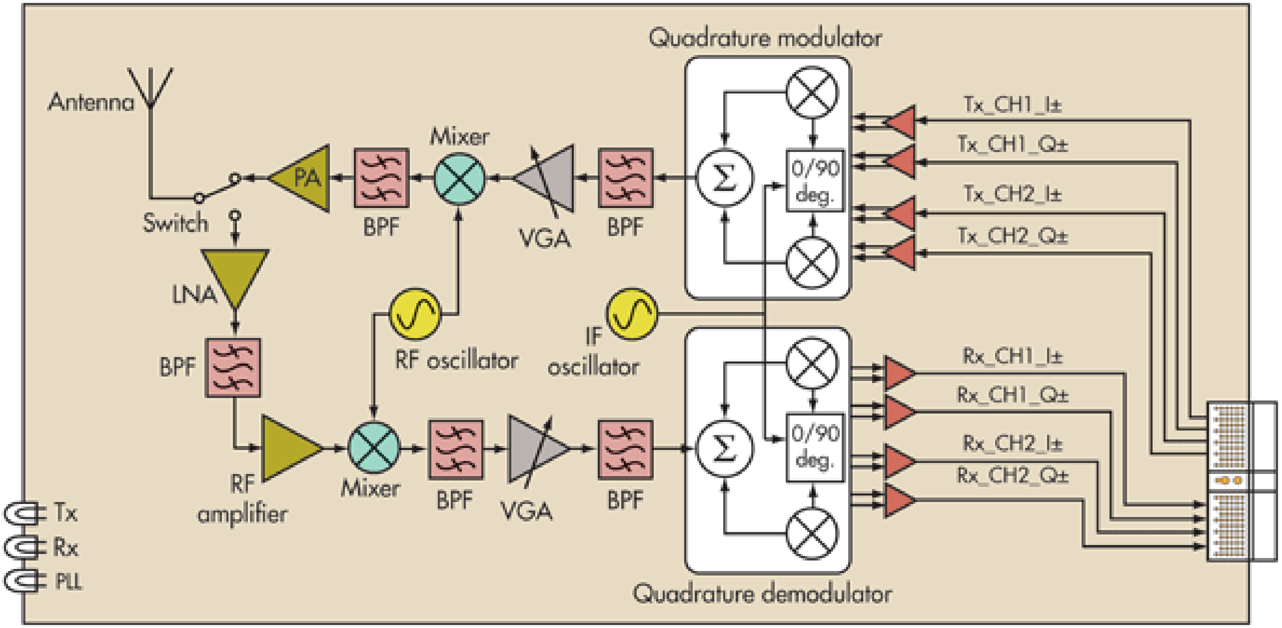
$$|\Delta| = |S_{11}S_{22} - S_{12}S_{21}| < 1$$

- In 1992, Edwards et. al. derived a new criterion that combines the k-Δ parameters into a test involving only a single parameter μ for unconditional stability. Thus, if μ>1, the device is unconditional stable. In addition, it can be said that large values of μ imply greater stability.

$$\mu = \frac{1 - |S_{11}|^2}{|S_{22} - \Delta S_{11}^*| + |S_{12}S_{21}|} > 1$$

The role of Mixers

- Mixers perform frequency translation by multiplying two waveforms (and possibly their harmonics).
- Three-port circuit requiring an LO reference input
  - Down-conversion in receivers, output frequency (IF) is the difference of two input frequencies (RF and LO)
  - Up-conversion in transmitters, output frequency (RF) is the sum of two input frequencies (IF and LO).



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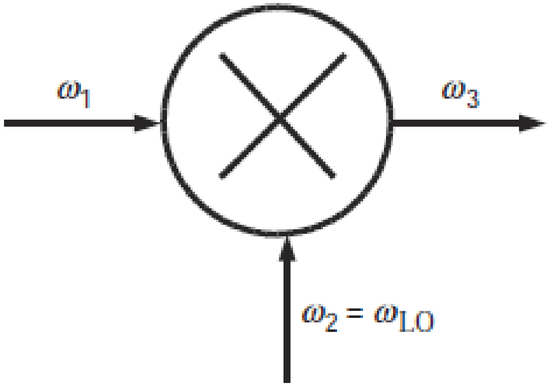
Basic operation (I)

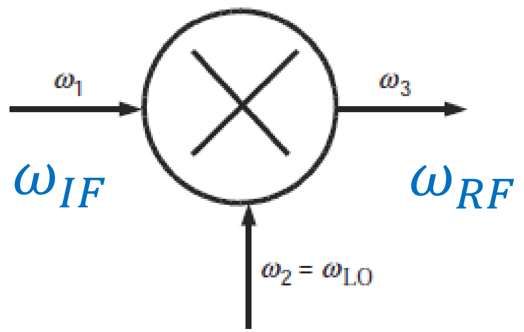
- The ideal behavior of the mixer as a multiplier.
- Signals are produced at the output at the sum and difference frequencies of the two inputs.
- In practice, the multiplier is implemented with a non-linear element such as a diode or transistor, or with a time-varying element.



$$A\cos(\omega_1 t) \times B\cos(\omega_2 t)$$
$$= \frac{AB}{2}[\cos(\omega_1 - \omega_2)t + \cos(\omega_1 + \omega_2)t]$$

Basic operation (II)

- **Upconverter mixer operation**
  - the mixer input signal is applied at the IF port $\omega_1=\omega_{IF}$, $\omega_2=\omega_{LO}$ and both the $\omega_{LO}-\omega_{IF}=\omega_{RF}$ and the $\omega_{LO}+\omega_{IF}=\omega_{RF}$ sidebands are generated at the RF (output) port.
  - The sum frequency $\omega_{RF}=\omega_{LO}+\omega_{IF}$ is known as the upper sideband (USB). The mixer also produces a lower sideband (LSB) at $\omega_{RF}=\omega_{LO}-\omega_{IF}$.
  - A bandpass filter, placed at the RF port, selects the desired signal ($\omega_{RF}=\omega_{LO}+\omega_{IF}$) and rejects any leakage from the IF and LO ports.



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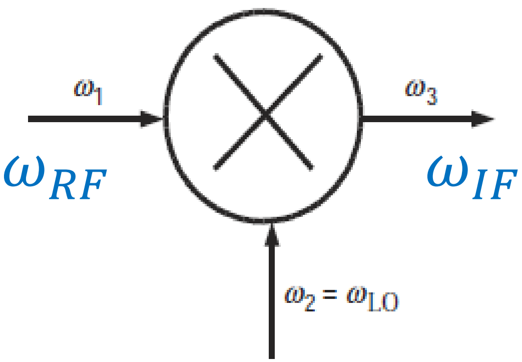
Basic operation (II)

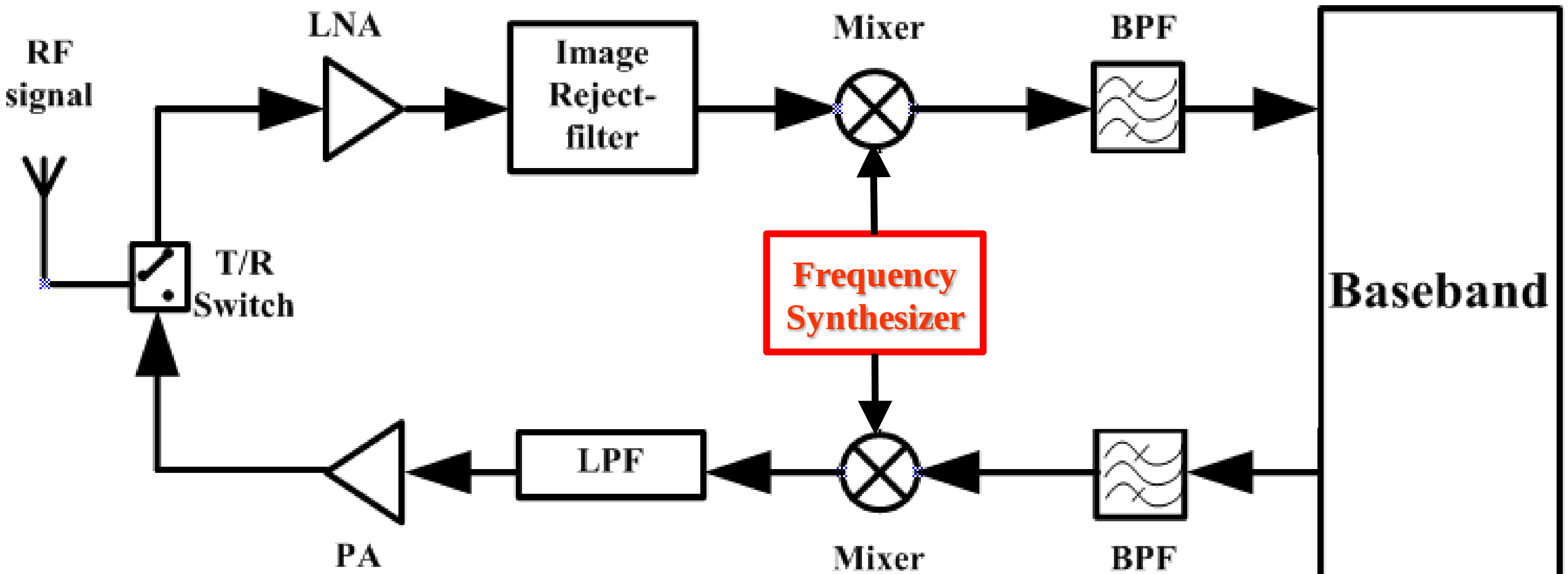
- **Downconverter mixer operation**
  - In a receiver, the input signal is applied to the RF port and $\omega_1 = \omega_{RF}$, where $\omega_{RF} = \omega_{LO} + \omega_{IF}$ or $\omega RF = \omega_{LO} - \omega_{IF}$.
  - The LO signal is applied at port 2, as in an upconverter, $\omega_2 = \omega_{LO}$, and signals both at $\omega_{RF} - \omega_{LO} = \omega_{IF}$ and $\omega_{LO} - \omega_{RF} = \omega_{IF}$ are collected at the IF port.



General Considerations

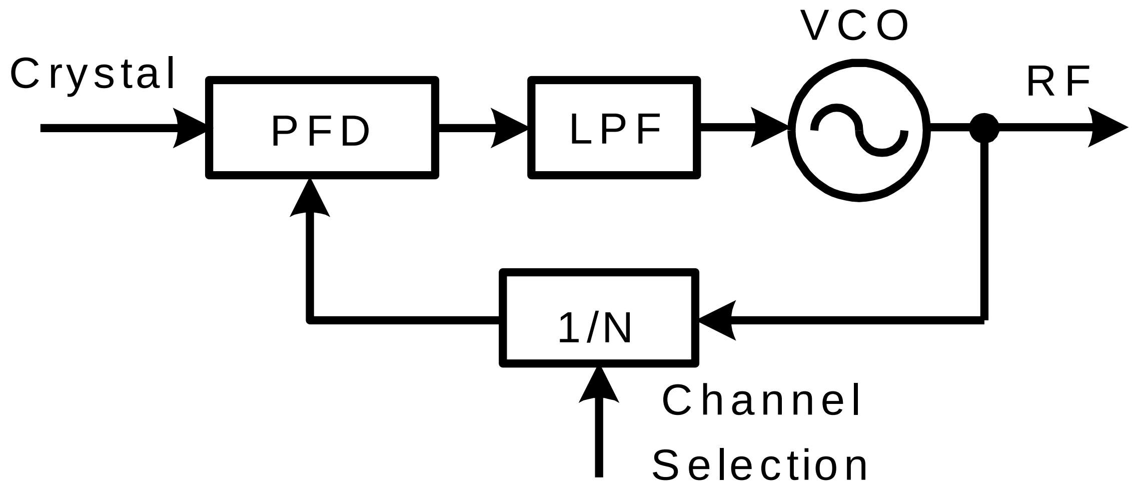
- Impedance matching
- Conversion gain
- Noise performance
- Isolation
- Linearity
- Spurious

Frequency Synthesizer in Wireless System

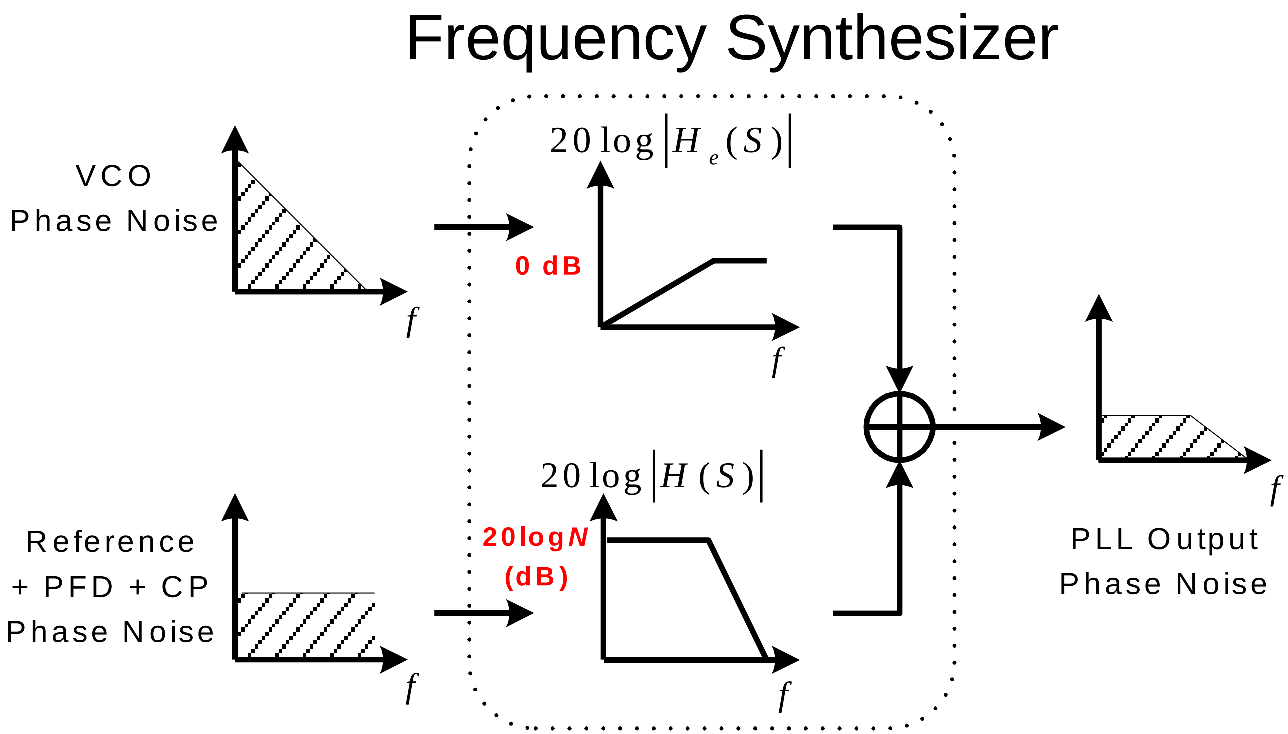


PLL-based Frequency Synthesizer

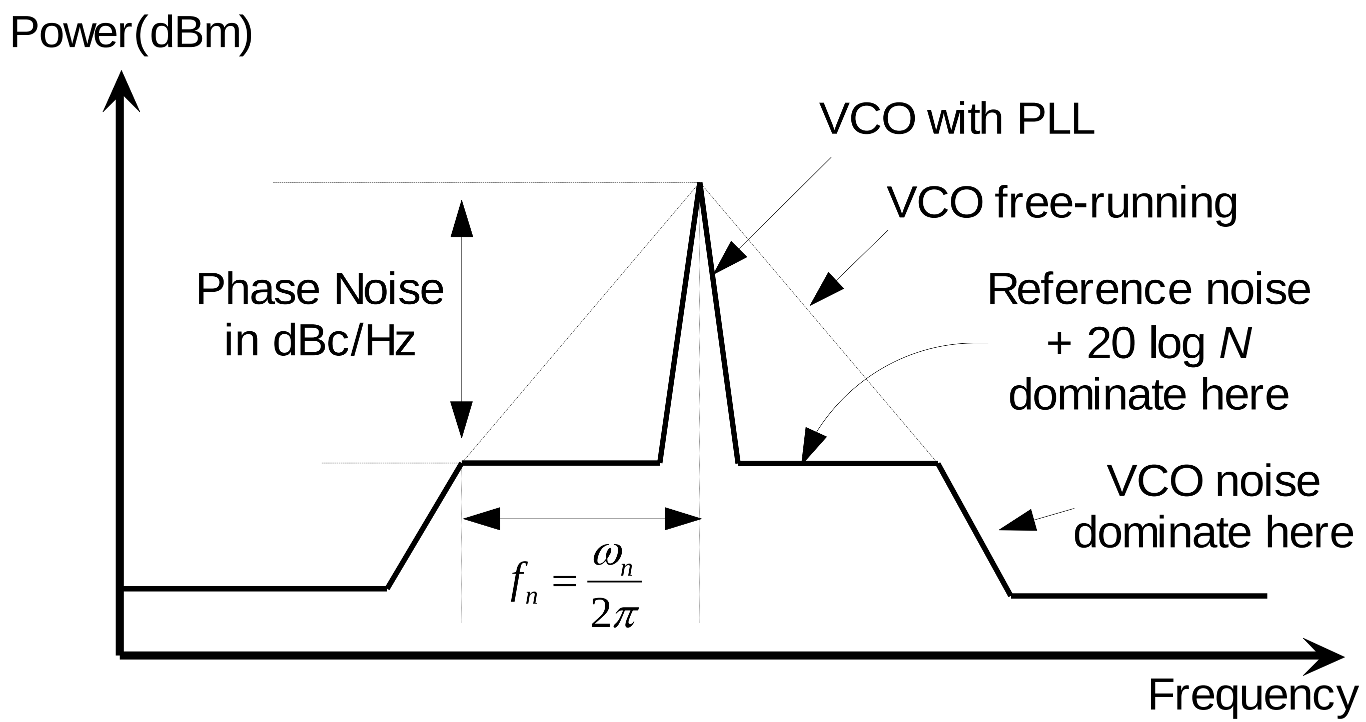
- Most frequency synthesizers are based on a phase-locked loop (PLL).
- The frequency and phase of the voltage-controlled oscillator (VCO) is locked by the PLL with reference to the crystal oscillator.
- The synthesized frequency can be adjusted by tuning the division value *N*.



Phase Noise Suppression



Phase Noise Properties and Measurement



$$\mathcal{L}(\Delta f) = 10\log\left(\frac{P_{sideband} @ \Delta f}{P_{carrier}} \frac{1}{\mathrm{RBW}}\right) = 10\log P_{sideband} @ \Delta f - 10\log P_{carrier} - 10\log \mathrm{RBW}\ \ (\mathrm{dBc/Hz})$$

RBW: the resolution BW of spectrum analyzer as measuring phase noise.

Fractional-*N*

Integer-*N*

$$f_{VCO} = N \times f_{ref}$$

$$f_{res} = \left[(N+1) - N \right] f_{ref}$$

$$= f_{ref} = \frac{f_{VCO}}{N}$$

Fractional-*N*

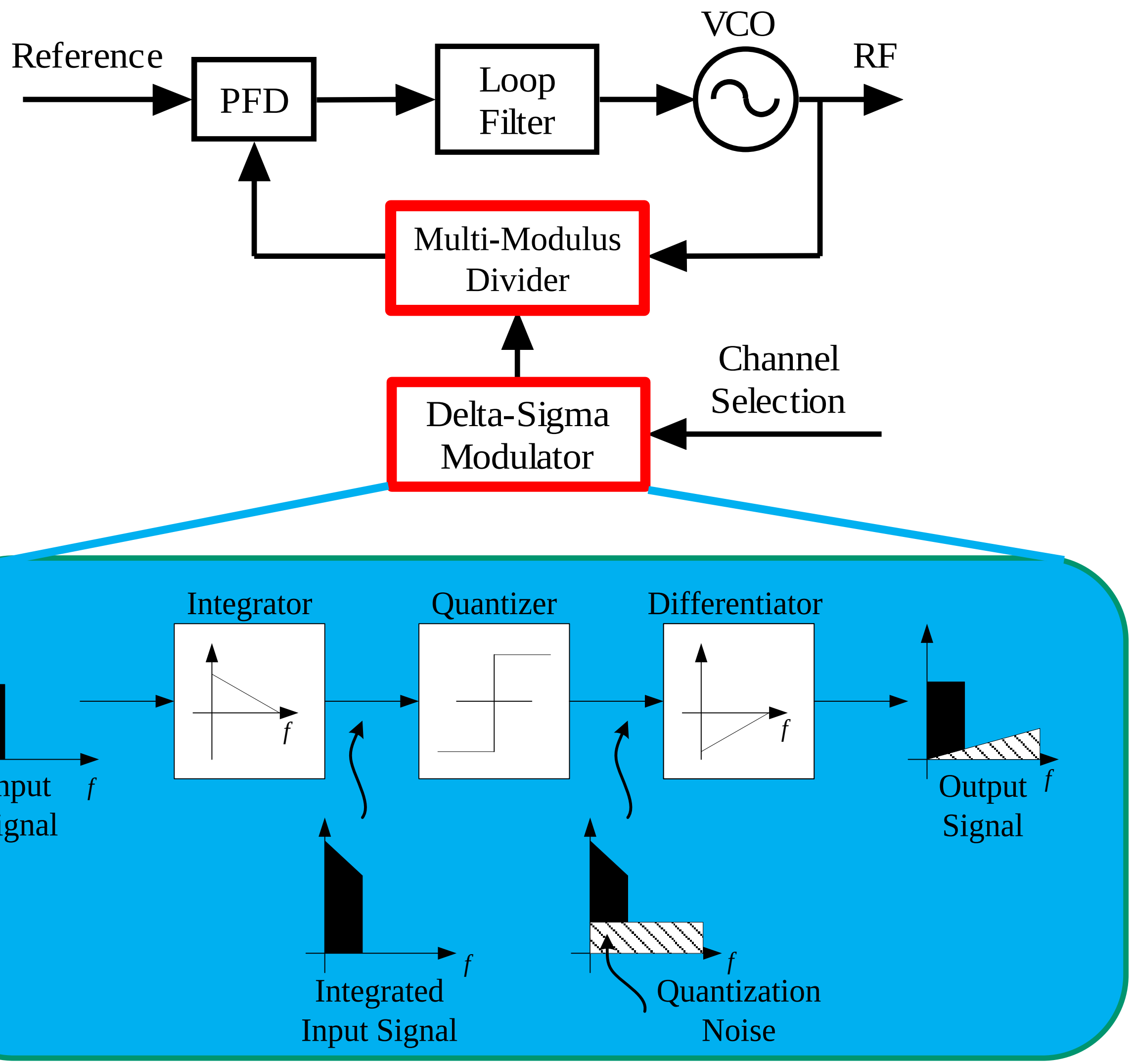
$$f_{VCO} = N \times f_{ref} = \left(P + \frac{A}{M} \right) \times f_{ref}$$

$$f_{res} = \left[\left(P + \frac{A+1}{M} \right) - \left(P + \frac{A}{M} \right) \right] f_{ref}$$

$$= \frac{f_{ref}}{M} = \frac{f_{VCO}}{M \cdot N}$$

The frequency resolution can be arbitrarily modified with *M*.

Fractional-*N* Frequency Synthesizer



Sub-6GHz/mmW共構之多輸入輸出射頻模組—系統效能分析與評估

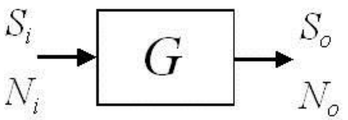
大綱

- Sub-6GHz/mmW共構之射頻接收機系統效能分析
- Sub-6GHz/mmW共構之射頻發射機系統效能分析
- Sub-6GHz/mmW共構之多輸入輸出天線系統效能分析
- Sub-6GHz/mmW共構之多輸入輸出射頻傳收機系統效能分析

射頻接收機系統效能

- 接收機鏈路增益
- 接收機雜音指數與接收靈敏度
- 接收機輸入三階折斷點與輸入信號動態範圍
- 接收機影像拒斥比（超外差式接收機）

線性系統雜訊指數



$$S_o = G \cdot S_i \qquad N_i = kT_0 B$$

$$N_o = GkT_0 B + GkT_e B; \quad T_e: \text{equivalent noise temperature of the system}$$

$$NR \text{ (noise ratio)} \equiv \frac{(S/N)_i}{(S/N)_o} = \frac{S_i}{S_o} \cdot \frac{N_o}{N_i} = \frac{1}{G} \frac{GkT_0 B + GkT_e B}{kT_0 B} = 1 + \frac{T_e}{T_0}$$

- 三級串接線性系統雜訊比

$$S_o = G_1 G_2 G_3 \cdot S_i \qquad N_i = kT_0 B$$

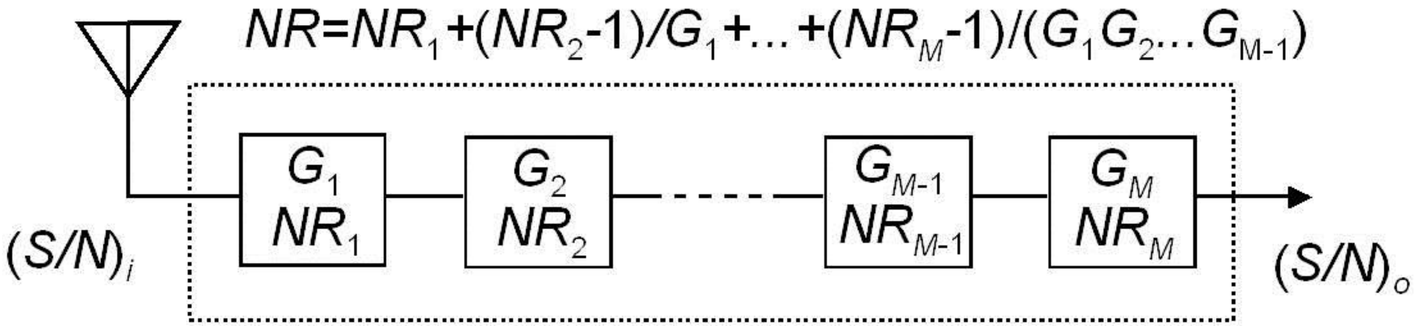
$$\begin{aligned} N_o &= G_1 G_2 G_3 kT_0 B + G_1 G_2 G_3 kT_{e1} B + G_2 G_3 kT_{e2} B + G_3 kT_{e3} B \\ &= G_1 G_2 G_3 N_i + G_1 G_2 G_3 (NR_1 - 1) N_i + G_2 G_3 (NR_2 - 1) N_i + G_3 (NR_3 - 1) N_i \\ &= G_1 G_2 G_3 NR_1 N_i + G_2 G_3 (NR_2 - 1) N_i + G_3 (NR_3 - 1) N_i \end{aligned}$$

$$NR_{total} = \frac{(S/N)_i}{(S/N)_o} = \frac{S_i}{S_o} \cdot \frac{N_o}{N_i} = \frac{1}{G_1 G_2 G_3} \frac{N_o}{N_i} = NR_1 + \frac{NR_2 - 1}{G_1} + \frac{NR_3 - 1}{G_1 G_2}$$

- *M*級串接線性系統雜訊指數

$$NR = NR_1 + \frac{NR_2 - 1}{G_1} + \cdots + \frac{NR_M - 1}{G_1 G_2 \cdots G_{M-1}} \qquad NF \text{ (noise figure)} = 10 \log_{10} NR$$

射頻接收系統雜訊指數與靈敏度



- 射頻接收系統輸出端之訊雜比$(S/N)_o$由整體雜訊比NR與輸入端之訊雜比$(S/N)_i$決定
- 解調器需有最低的訊雜比確保其誤碼率可保持在一定品質以上，由此可訂出此射頻接收系統之靈敏度（Sensitivity）

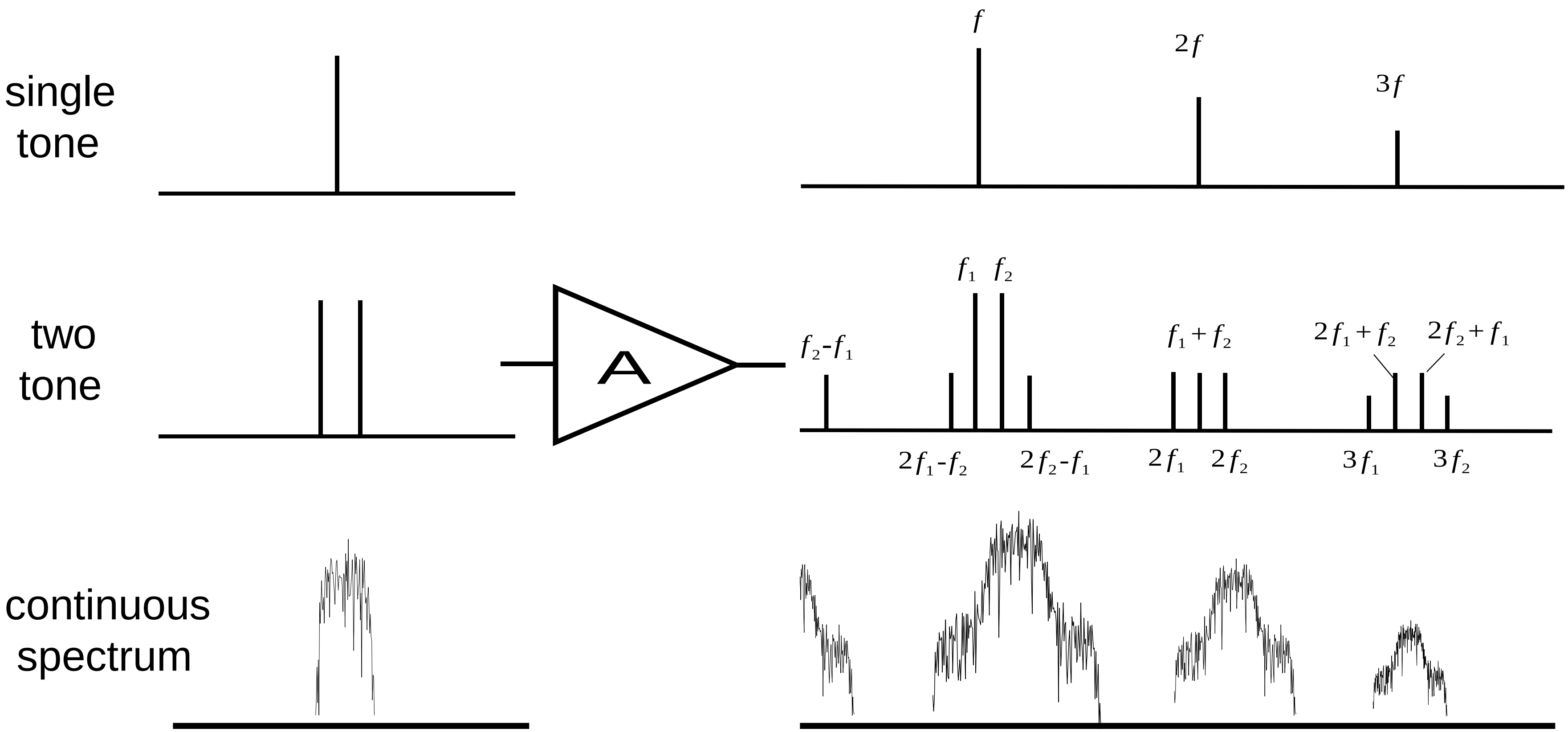
$$S_i = NR \cdot (S/N)_{o,\min} \cdot N_i, \quad N_i = kTB$$

$$S_{i,dBm} = NF + (S/N)_{o,\min,dB} + 10\log_{10} B - 174 \Big|_{T=290^\circ K}$$

- 射頻接收機第一級電路之雜訊指數具決定性影響

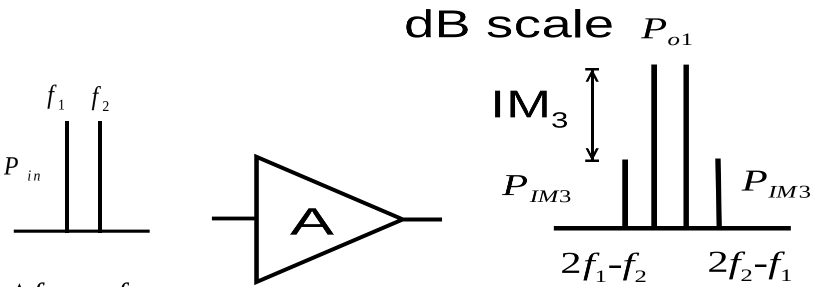
非線性失真

- The nonlinear effects yield signal impairments, including harmonic generation, intermodulation, and spectral regrowth.



教育部5G行動寬頻人才培育計畫
5G天線與射頻技術聯盟中心

三階交互調變失真與動態範圍



$\Delta f \ll f_1$

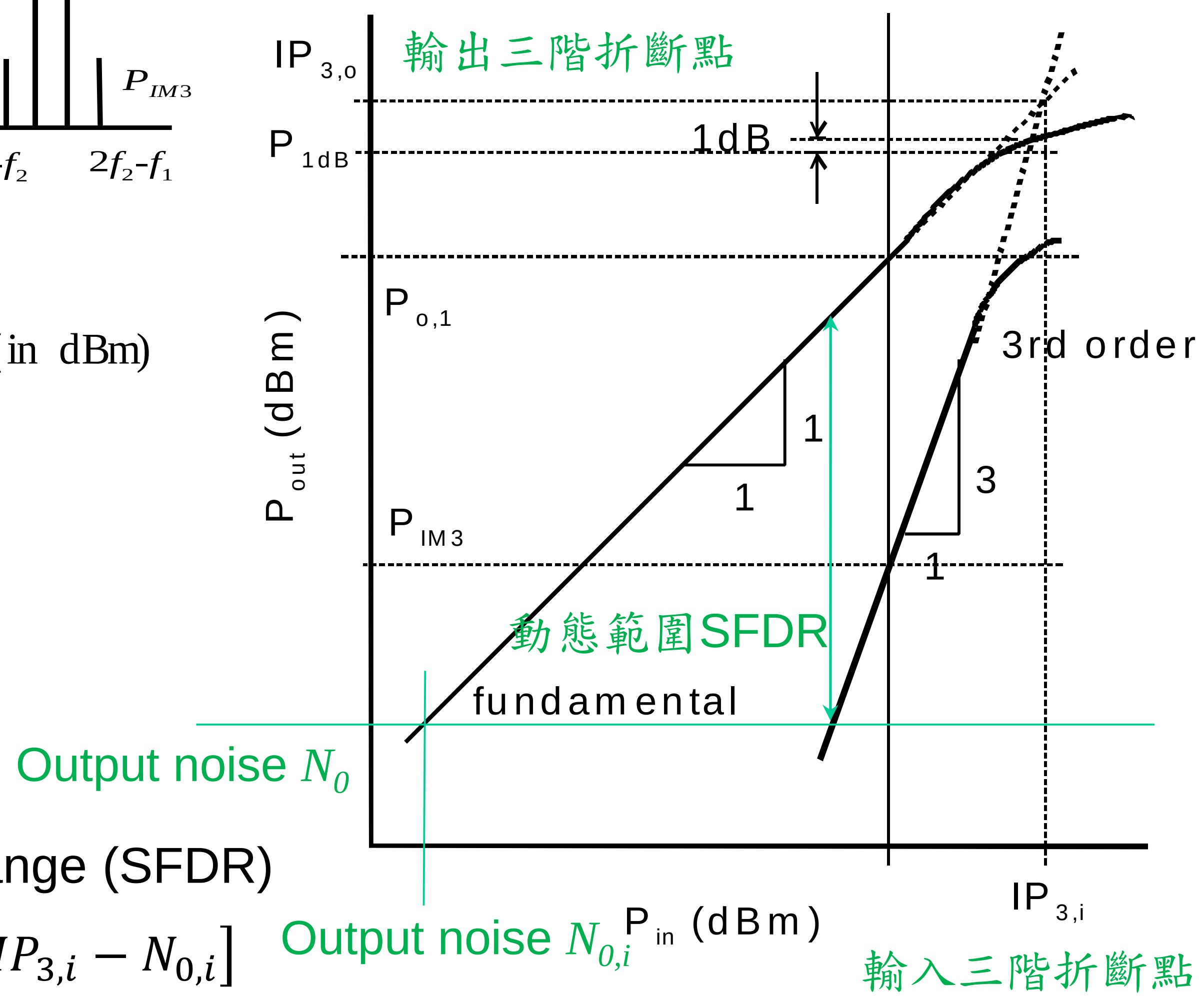
$\Delta f \ll f_2$

$$(IP_{3,o} - P_{IM,3})/3 = IP_{3,o} - P_{o,1} \text{ (in dBm)}$$

$$2P_{o,1} + \underbrace{(P_{o,1} - P_{IM,3})}_{IM_3\text{ (dBc)}} = 2IP_{3,o}$$

$$\Rightarrow IP_{3,o} = \underbrace{P_{o,1}}_{G+P_{in}} + 0.5\underbrace{IM_3}_{\text{specified by designer}}$$

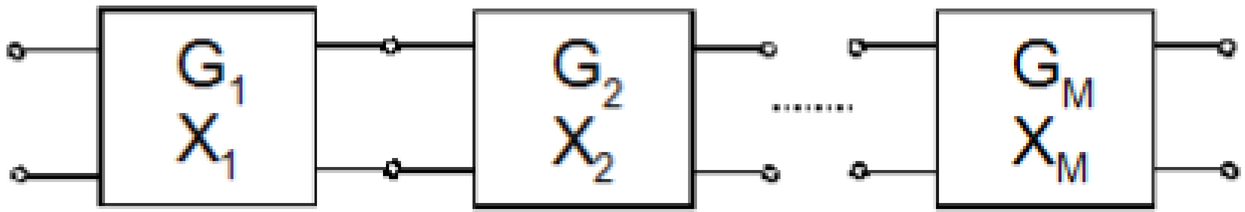
$$\text{or} \quad P_{IM_3} = 3P_{o,1} - 2IP_{3,o}$$



Spurious Free Dynamic Range (SFDR)

$$\text{SFDR} = \frac{2}{3}\left[IP_{3,o} - N_0\right] = \frac{2}{3}\left[IP_{3,i} - N_{0,i}\right]$$

射頻接收系統三階折斷點



G_n : *gain*

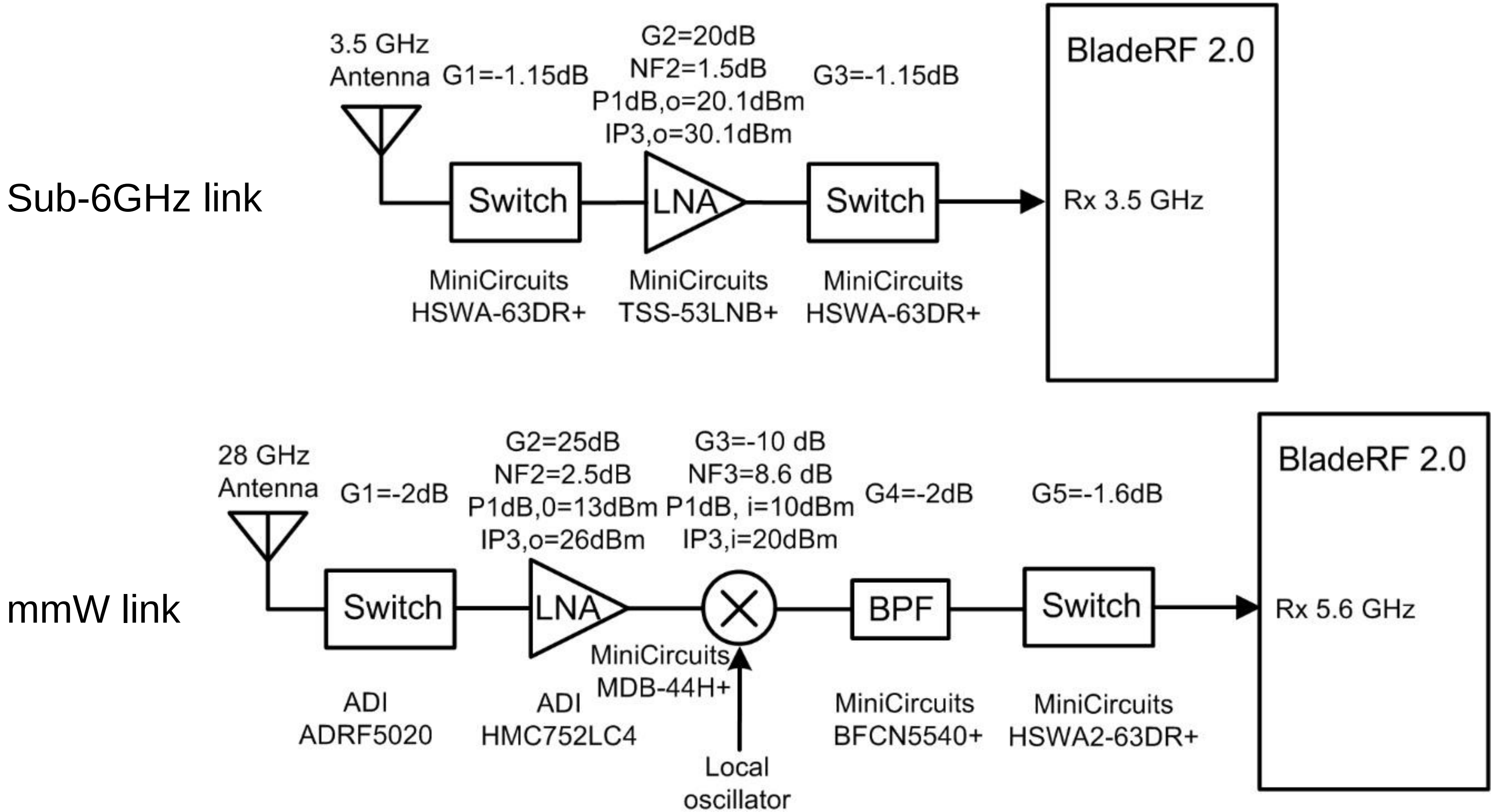
X_n : I/P or O/P IP3

$$X_{total}^{\min} = \left(\sum_{n=1}^{M} \frac{1}{X_n G_n} \right)^{-1}$$

$$X_{total}^{\max} = \left(\sum_{n=1}^{M} \frac{1}{X_n^2 G_n^2} - 2 \sum_{\substack{n=2 \\ n>k}}^{M} \sum_{k=1}^{M} \frac{1}{X_n X_k G_n G_k} \right)^{-1/2}$$

Ref: N. G. Kanaglekar, R. E. McIntosh, and W. E. Bryant, “Analysis of two-tone, third-order distortion in cascaded two-ports,” *IEEE Trans. Microw. Theory Techn.*, vol. MTT-36, pp. 701-705, Apr. 1988.

Sub-6GHz/mmW共構之射頻接收機系統



- Performance evaluation by approximated formulas or CAD tool

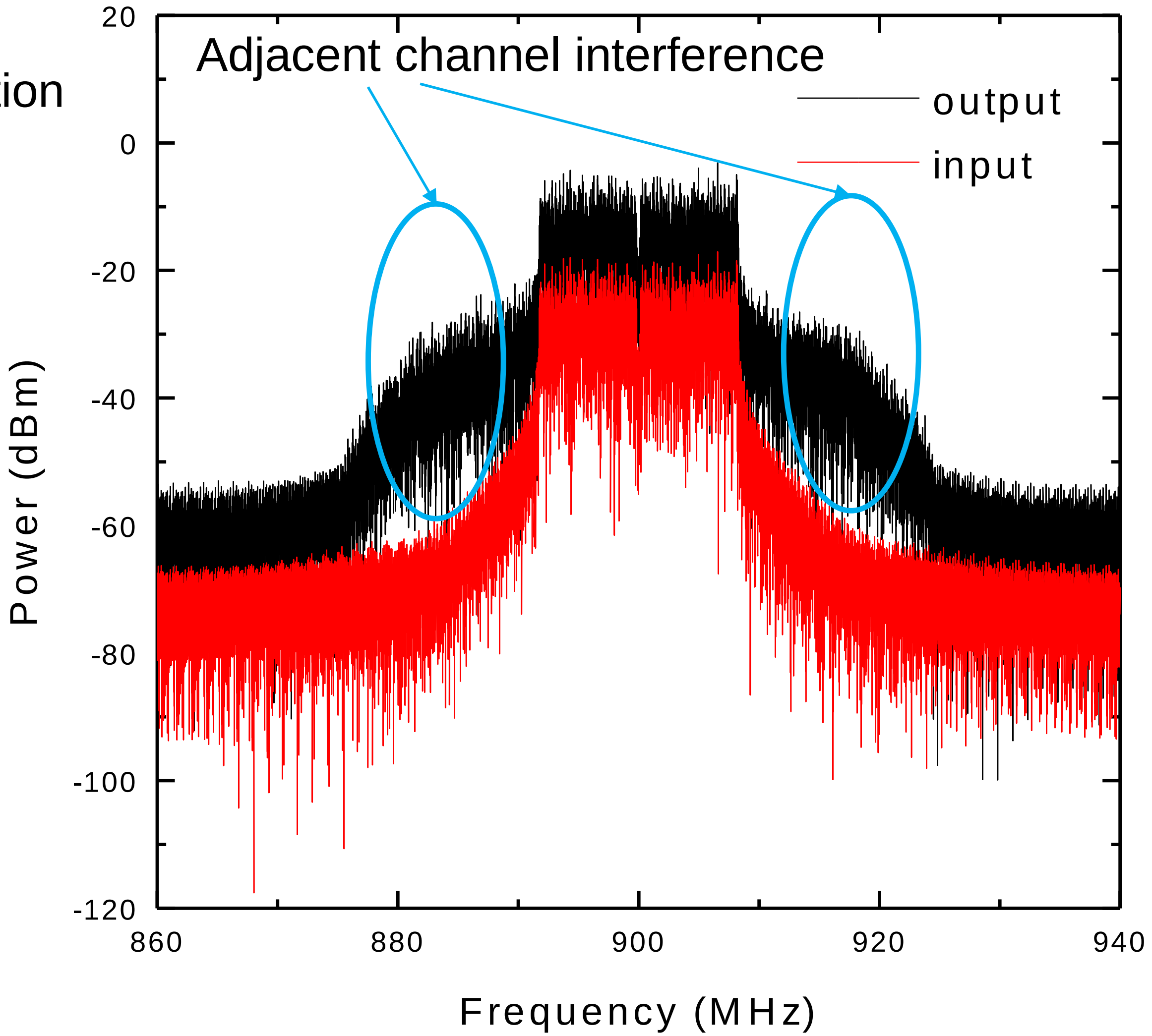
Sub-6GHz/mmW射頻發射機系統效能

- 發射機鏈路增益
- 發射機輸出1dB增益壓縮點、三階折斷點與輸出功率
- 發射機信號失真度
- 發射機輸出諧波與雜波

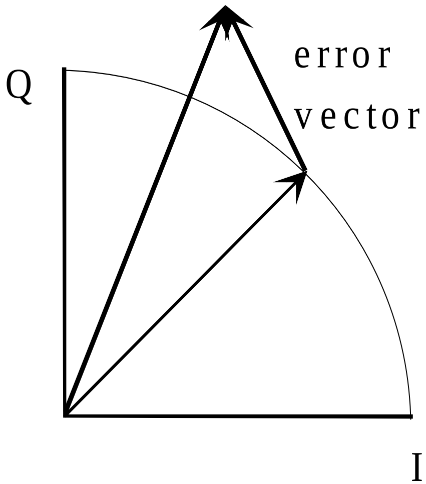
發射機輸出信號鄰頻干擾

Power amplifier nonlinear distortion

OFDM/16QAM modulation

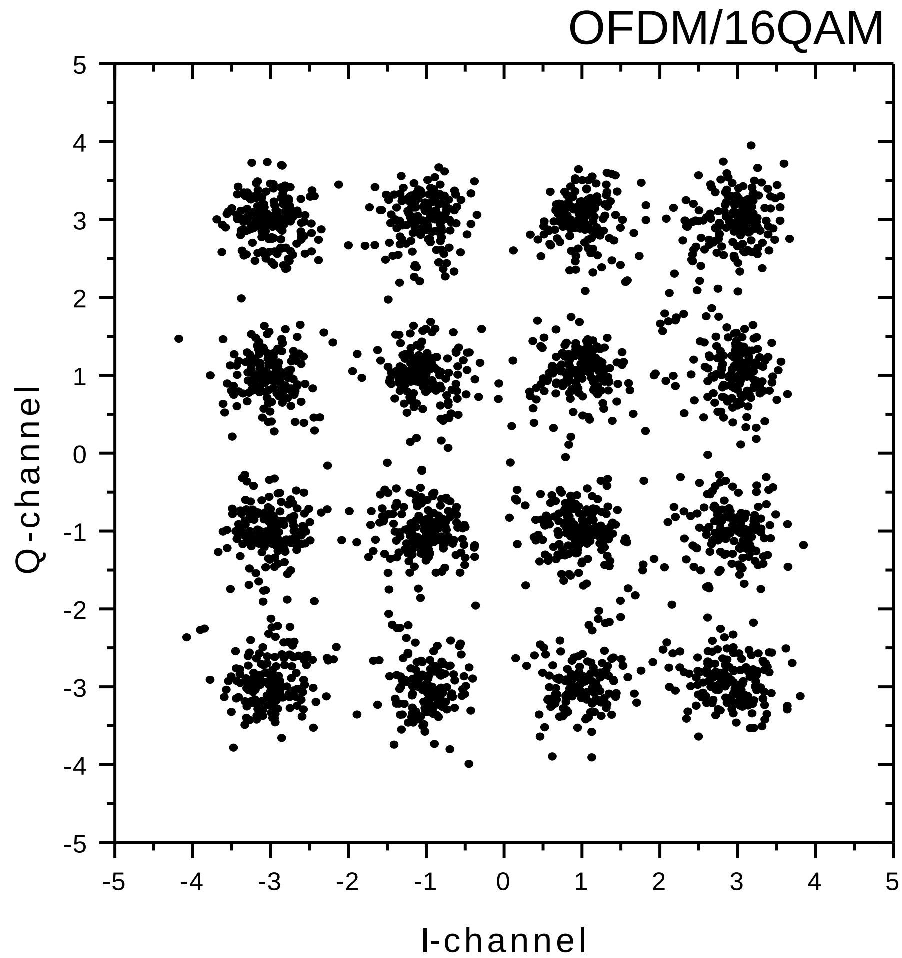


發射機輸出信號失真度



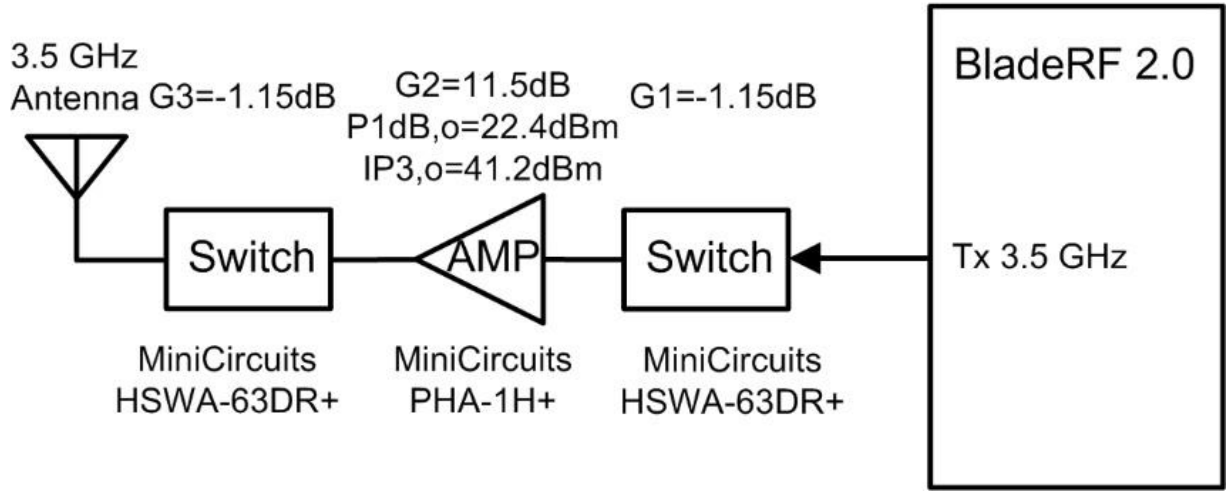
Error Vector Magnitude

$$EVM_{RMS} = \sqrt{\frac{\sum_{j=1}^{L_s}\left[\sum_{k=1}^{52}\left(C_k^{distorted} - C_k^{ideal}\right)^2\right]}{52 \times L_s \times P_0}}$$

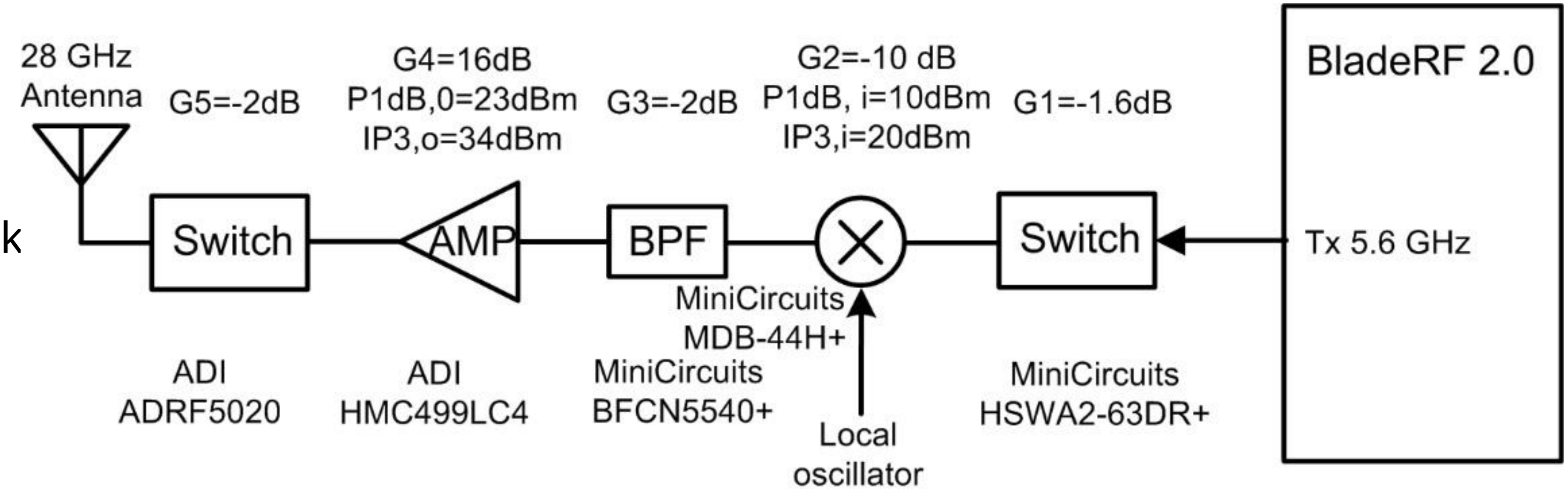


Sub-6GHz/mmW共構之射頻發射機系統

Sub-6GHz link



mmW link



- Performance evaluation by approximated formulas or CAD tool

Sub-6GHz/mmW多天線輸入輸出系統

- 目的
  - 增加天線系統等效全向輻射功率
  - 提升傳輸信號品質
  - 增加資訊傳輸率
- 實現方式
  - 平面式天線搭配嵌入式射頻模組
  - 可調整每個天線信號的大小與相角
  - 可執行不同功能
    - ✓傳輸分集（Transmit diversity）
    - ✓空間複用（Spatial multiplexing）
    - ✓波束成型（Beamforming）

波束成型—相位陣列天線輻射場型關係

- 陣列天線輻射場型可表為天線元件因子（antenna element factor）乘以陣列因子（array factor）

$$F(\theta,\varphi) = F_{array}(\theta,\varphi) \cdot F_{element}(\theta,\varphi)$$

- 若陣列天線為排列在*z*-軸等間距*d*的2*N*+1個線性陣列

$$F_{array}(\theta) = \sum_{n=-N}^{N} I_n e^{jkndcos\theta}, \; k = \omega\sqrt{\mu_0\epsilon_0},$$

- 若饋入天線之電流大小相同、相移為等差相位

$$I_n = I_0 e^{-jn\alpha_z}, F_{array}(\theta) = \sum_{n=-N}^{N} e^{jn(kdcos\theta-\alpha_z)}$$

- 調整α_z即可調整主波辦輻射角度θ_0

$$kdcos\theta_0 = \alpha_z, \; \theta_0 = cos^{-1}(\frac{\alpha_z\lambda}{2\pi d})$$

相位陣列天線輻射場型實際狀況

- 陣列因子較易計算掌控
- 陣列天線整體輻射場型須包含天線元件因子，天線指向性越高，對整體場型影響越大
- 天線間之耦合效應亦會造成整體輻射場型變異
- 陣列天線之振幅調整（可調增益放大器或衰減器）以及相移器相移誤差造成陣列因子偏移
- 振幅調整會造成額外相移，相移調整也會造成各額外損耗，皆影響陣列因子結果

Sub-6GHz/mmW多天線輸入輸出系統效能

- 波束成型效能
  - 可掃描最大角度
  - 主波瓣掃制最大角度時，主波瓣波束寬度變化量
  - 主波瓣掃制最大角度時，旁波瓣變化量

各種波束成型設計架構-I

- 射頻端相移
- 本地振盪端相移
- 中頻／基頻端類比式相移
- 數位式相移
- 混合波束成型架構

波束掃描對射頻傳收機系統效能影響

- 不同相位移時，可能造成射頻鏈路衰減量不同
- 接收雜訊指數可能在不同相位移時產生不同值
- 發射功率可能在不同相位移時產生不同結果
- 發射信號失真度可能在不同相位移時有不同失真量
- 數位式相移對射頻傳收機效能影響量最低

Sub-6GHz/mmW共構之多輸入輸出射頻模組－與天線陣列之整合技術

大綱

- 半導體封裝與印刷電路板整合技術概觀
- 應用封裝與印刷電路板製程於天線陣列設計
- 應用封裝與印刷電路板製程於天線陣列與射頻模組整合
- 電源完整性、信號完整性、電磁干擾與散熱問題

半導體封裝技術概觀

- Embedded Wafer-Level Ball grid array (eWLB)
- Redistributed Chip Package (RCP)
- Integrated Fan-out Wafer Level Package (InFO-WLP)
- Embedded Wafer Level Package-on-Package (eWLB-PoP)

射頻模組與天線陣列整合設計

- 射頻模組與天線陣列高密度整合下之設計挑戰
  - 高密度整合下電源分配網路之電源完整性問題
  - 高密度整合下類比/數位信號傳輸之信號完整性問題
  - 高密度整合之電磁干擾問題
  - 高密度整合之散熱與可靠度問題
- 可用之分析模擬技術
  - 電路/電磁/天線整合模擬分析技術
  - 熱/電/應力之多重物理模擬分析技術