RECEIVING SYSTEMS
FOR
RADIO ASTRONOMY

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Introduction

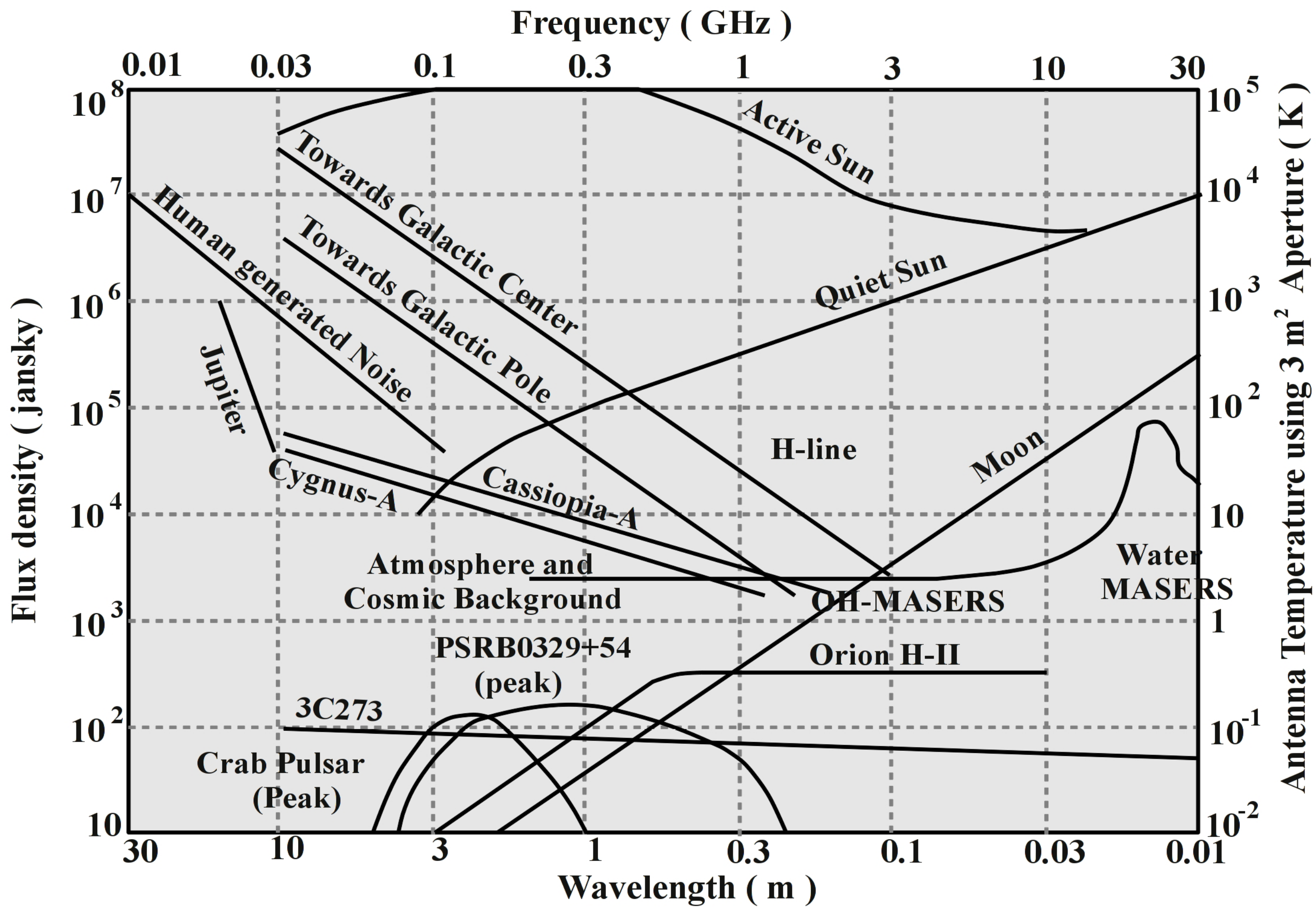
In this lecture we shall study the basic receiving systems used in Radio Astronomy.

In any radio telescope, antennas are followed by receiving systems which further process the signals. Except for transient emissions from Sun and Jupiter, in general the signals are very weak (micro and milli-jansky). Hence, the system requirements are:

(*i*) High sensitivity.
(*ii*) High gain.
(*iii*) High stability.

The signal strength and corresponding antenna temperature plays a major role in determining the receiver requirements. Therefore, we first look at some of the antenna temperatures produced on Earth surface by a few astronomincal sources with known brightness.

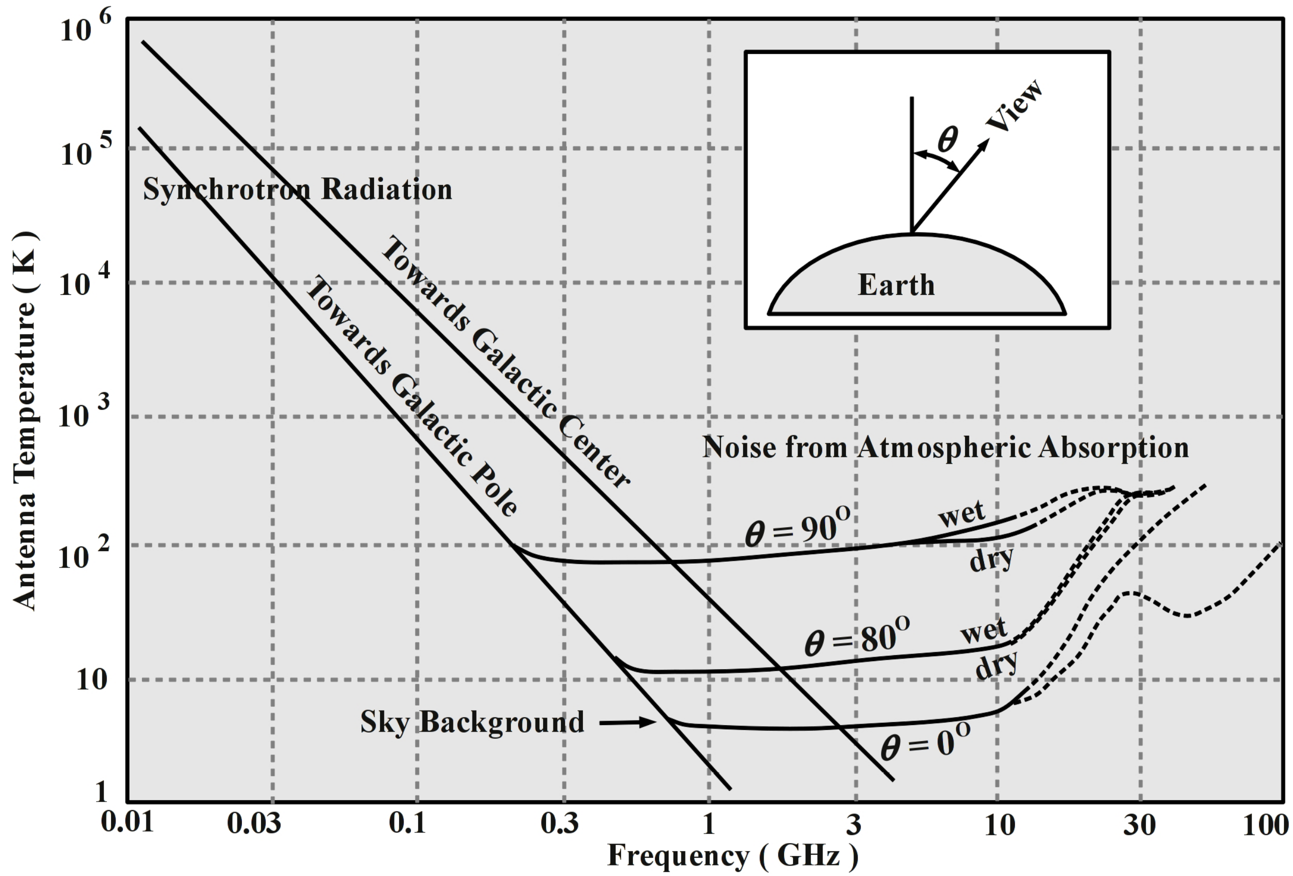
Introduction



As shown above, the dynamic range of radio emissions and spectra of the astronomical sources are very large. The corresponding antenna temperatures using a $3 m^2$ aperture antenna are also shown. Due to different spectral ranges of different sources, a single frequency-band antenna is not able to accommodate them. Thus, different set of antennas having different receiving systems can be used.

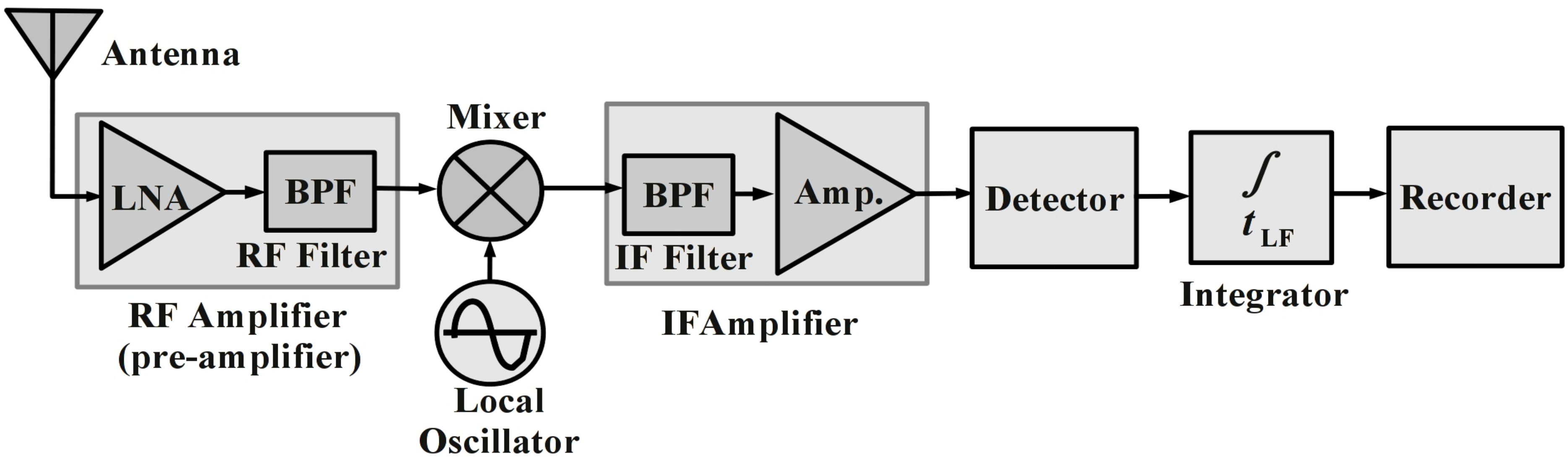
Atmosphere on Receiving Sensitivity

Variations of antenna temperature of the sky as a function of frequency. The beam angle of the antenna assumed to be less than a few degrees having 100% beam efficiency.



However sensitive you make a radio telescope, its capabilities are limited by Earth atmosphere. Depending on viewing direction θ of the telescope (w.r.t. local zenith), atmospheric noise may limit the threshold of observation for frequencies above 0.2 GHz as shown. Below this frequency, noise from galactic center and galactic pole dominates. The atmospheric noise also depends on the weather conditions (wet or dry) above 3 GHz. The best one can achieve is by observing any source at the zenith.

Parameters of Basic Rec. System (Radiometer)



Important parameters of a Receiver

1. **RF Center Frequency** ν_0: Center frequency of the RF band chosen for observation (same as the center frequency of the RF filter). Unit is Hz.

2. **Gain** G_{Rcv}: It is dimensionless. $$G_{\mathrm{Rcv}} = \frac{\text{Power output to recorder}}{\text{Power input to receiver}}$$

3. **Band-width** $\Delta\nu_{HF}$: It is the rectangular observation band-width of the receiver. Since RF and IF band-widths are identical, $\Delta\nu_{HF}$ represent both. Unit is Hz.

Parameters of Basic Rec. System (Radiometer)

4. <u>Integration Time</u> τ_{LF}: The detector output keeps fluctuating with time. To smooth out the large scale fluctuations before recording, integration is applied over a fixed period of time. Measured is sec. It also improves sensitivity.

5. <u>Receiver Temperature</u> T_R: It is the effective input noise temperature seen at the input terminals of a receiver. ***Over a bandwidth*** $\Delta\nu_{HF}$***, the temperature at which a resistor (equivalent to input resistance of receiver) generates same amount of noise power as that of the receiver is defined as receiver temperature*** T_R. It is measured in K. The noise power W_R (watt) of a receiver is given as: $W_R = k\,T_R\,\Delta\nu_{\text{HF}}$

T_R is related to the ***noise factor*** F of the receiver as: $T_R = (F-1)\times T_{R(\text{Phy})}$

Here, $T_{R(\text{Phy})}$ is the physical temperature of receiver in K. The noise factor when converted to dB is known as ***noise figure***. Since receivers are composed of cascaded stages, T_R can be expressed as a function of noise contribution from each stage. $T_R = \frac{1}{\epsilon_T}\left(T_1 + \frac{T_2}{G_1} + \frac{T_3}{G_1\,G_2} + \ldots\right)$

T_1, T_2, ... are respectively the individual temperatures generated at inputs of the 1[st], 2_{nd}, ... stages, and ε_T is the efficiency of the transmission line joining the receiver with the antenna.

Radiometer: Sensitivity and Stability

1. **Sensitivity ΔT_{min}:** It is the smallest change in antenna temperature ΔT_{min} which the instrument can detect successfully. Even for a perfectly stable receiver, higher T_{Sys} can reduce the sensitivity. It is given as:

$$\Delta T_{\min} = K_s \, \frac{T_{Sys}}{\sqrt{\tau_{LF} \, \Delta\nu_{\mathrm{HF}}}}$$

Here, $\Delta\nu_{HF}$ is the equivalent rectangular RF pass-band (same as IF), τ_{LF} is the integration time and K_s is the factor of sensitivity of the radiometer.

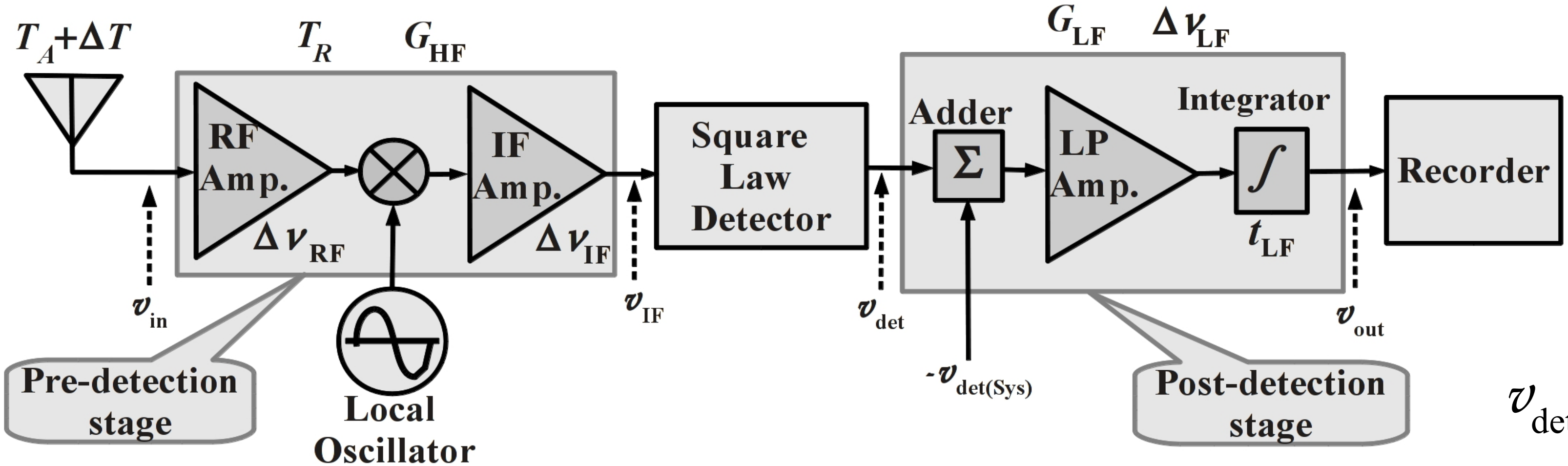
2. Stability: Over the period of observation, variation of gain, band-width and receiver noise temperature must be extremely minimized, though it may not be possible to make the system 100% stable. However, good LNA designs and use of switched radiometers may solve many of these problems. We shall discuss more about them in the following sections.

Types of Radiometer

Basic principles of different radiometers used in radio astronomy are discussed here. Some could be of historical importance. The objective for these designs had been to improve the sensitivity and stability. The following radiometers will be discussed:

- 1. Total Power Receiver (Direct Radiometer)
- 2. Dicke Receiver
- 3. Gain Modulated Dicke Receiver
- 4. Null Balancing Dicke Receiver
- 5. Graham's Receiver
- 6. Correlation Receiver
- 7. Additive Interferometer Receiver
- 8. Multiplicative Interferometer Receiver
- 9. Phase Switched Receiver

Total Power Receiver (Direct Radiometer)-I



$\Delta\nu_{RF} = \Delta\nu_{IF}$

Detector voltage

$v_{det(Src)}$ - source

$v_{det(Sys)}$ - cold sky

$v_{det} = v_{det(Sys)} + v_{det(Src)}$

Measures total noise power from antenna. The pre-detection stage generates a temperature T_R near the antenna terminal. T_A is antenna temperature of cold sky. ΔT is incremental temperature (antenna moved from cold sky to source). T_{Sys} is system temperature (antenna in cold sky) and k is Boltzmann constant.

IF power output is

$$W_{HF} = G_{HF}\,(T_{Sys} + \Delta T)\,k\,\Delta\nu_{RF} \quad \ldots (1)$$

Detector output voltage

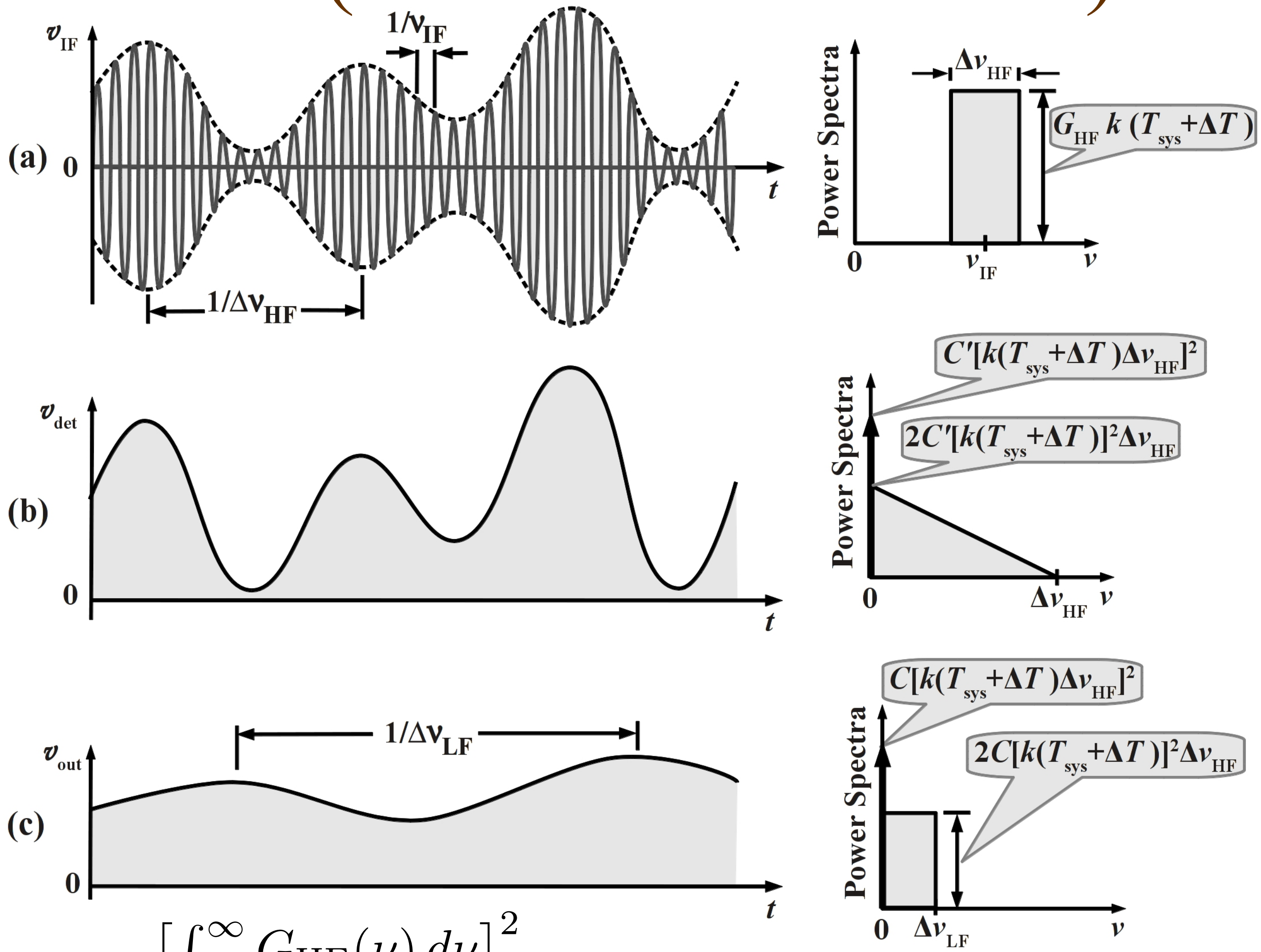
$$v_{det} = K_{det}\,G_{HF}\,(T_{Sys} + \Delta T)\,k\,\Delta\nu_{RF} = v_{det(Sys)} + v_{det(Src)} \quad \ldots (2)$$

where, $v_{det(Sys)} = T_{Sys}\,(K_{det}\,G_{HF}\,k\,\Delta\nu_{RF})$

and $v_{det(Src)} = \Delta T\,(K_{det}\,G_{HF}\,k\,\Delta\nu_{RF})$

Note: Generally, $v_{det(Src)} << v_{det(Sys)}$. To remove $v_{det(Sys)}$, $-v_{det(Sys)}$ is added in post detection. When pointed to cold sky, $v_{out} = 0$. When pointed to source (stronger than cold sky), $v_{out} = K_{LF}\,\Delta T\,(K_{det}\,G_{HF}\,k\,\Delta\nu_{RF}) \quad \ldots (3)$ K_{LF} – post detector gain

1. Total Power Receiver (Direct Radiometer)-II



Pseudo Waves & Spectra

(a) IF output voltage and power spectra.

(b) Detector output voltage and power spectra.

(c) Output voltage to the recorder and power spectra.

The dc component of a square law detector is proportional to the square of input ac voltage (power).

$G_{HF}(\nu)$ & $G_{LF}(0)$ - Pre and post detection (dc) gains.

Pre-detection bandwidth: $$\Delta\nu_{HF} = \frac{\left[\int_0^\infty G_{HF}(\nu)\,d\nu\right]^2}{\int_0^\infty \left[G_{HF}(\nu)\right]^2 d\nu} \quad \ldots (4)$$

Post-detection bandwidth: $$\Delta\nu_{LF} = \frac{\int_0^\infty G_{LF}(\nu)\,d\nu}{G_{LF}(0)} \quad \ldots (5)$$

Sensitivity (minimum detectable temperature): $$\Delta T_{min} = T_{Sys}\sqrt{\frac{2\,\Delta\nu_{LF}}{\Delta\nu_{HF}}} = (T_A + T_R)\sqrt{\frac{2\,\Delta\nu_{LF}}{\Delta\nu_{HF}}} \quad \ldots (6)$$

1. Total Power Receiver (Direct Radiometer)-III

If integration time is τ_{LF}, we may write the post detector bandwidth as:

$$\Delta\nu_{\rm LF} = \frac{1}{2\,\tau_{\rm LF}} \quad \ldots (7)$$

Substituting (7) in (6) we get:

$$\Delta T_{\rm min} = \frac{T_{Sys}}{\sqrt{\Delta\nu_{\rm HF}\,\tau_{\rm LF}}} \quad \ldots (8)$$

Effective integration time τ_{LF} for a smoothing filter can be obtained using (5) and (7) as:

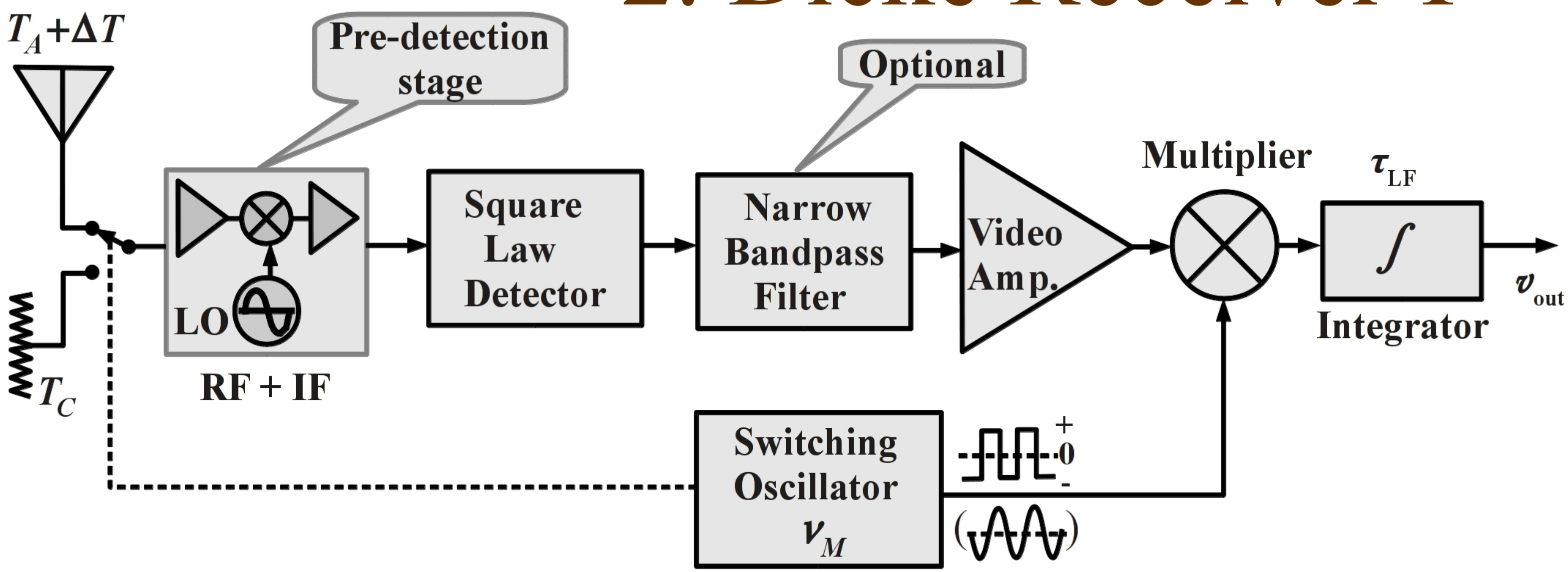
$$\tau_{\rm LF} = \frac{G_{\rm LF}(0)}{2\int_0^\infty G_{\rm LF}(\nu)\,d\nu} \quad \ldots (9)$$

Fluctuations in v_{out} due to amplifier gain variations are independent of the system noise fluctuations. If ΔG is the effective variations in G_{HF}, then the actual sensitivity is:

$$\Delta T_{\rm min} = T_{Sys}\sqrt{\frac{1}{\Delta\nu_{\rm HF}\,\tau_{\rm LF}} + \left(\frac{\Delta G}{G_{\rm HF}}\right)^2} \quad \ldots (10)$$

Post detection gain variations does not effect the sensitivity if dc compensation is used. Band-width and noise temperature variations of the receiver reduces the sensitivity as it happens with gain instability. An automatic gain control or AGC is not used here, for it can vary the receiver band-width and noise figure thereby introducing spurious noise fluctuations.

2. Dicke Receiver-I



G_{HF} - gain of pre-detection stage

ΔG - gain variations in pre-detection stage

RF input is switched between a resistive load and antenna at a frequency ν_M (10 to 10^4 Hz, 50% duty cycle). Multiplier changes sign of signals at frequency ν_M. After integration only the difference signals remain. If resistor temperature T_C is adjusted same as antenna temperature of cold sky, integrator output will depend only on incremental temperature ΔT. Thus for cold sky output becomes zero, and for source output goes positive.

Temperature variation ΔT_G at output due to gain variation ΔG is:

$$\Delta T_G = (T_A - T_C)\,\frac{\Delta G}{G_{\mathrm{HF}}} \quad \ldots (1)$$

If $T_A = T_C$, output voltage fluctuations due to gain instability disappears when antenna is pointed to cold sky. However, if $\Delta T \neq 0$ (when pointed to hot source) system gain changes do affect the output voltage.

2. Dicke Receiver-II

Due to switching, signal arrives for half observation time and sensitivity reduces by √2. Subtraction of noise signals further degrades it by √2.

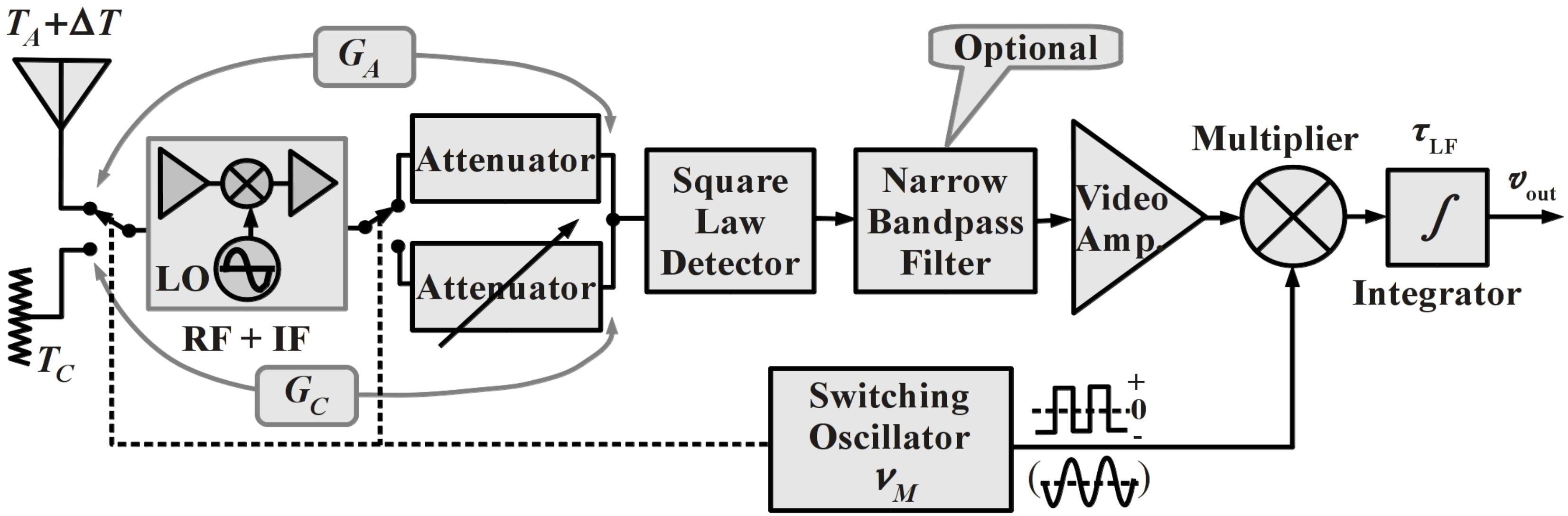
Sensitivity of a Dicke receiver is: $$\Delta T_{\min} = 2\frac{T_{Sys}}{\sqrt{\Delta\nu_{\mathrm{HF}}\,\tau_{\mathrm{LF}}}} \quad \ldots (2)$$

The switching waveform at multiplier input can be a square wave (preferable) or a sine wave. If square wave, video amplifier band-width should be about 10 ν_{M} to accommodate all the important harmonics. If video band-width is restricted to a few Hz centered at ν_{M}, with the aid of the optional narrow bandpass filter shown, some reduction in sensitivity takes place. The first harmonic amplitude of a square wave is 4/π times its amplitude. Hence its effective value is 4/√(2π). Thus, sensitivity in this case is obtained by multiplying (2) with √(2π)/4:

$$\Delta T_{\min} = \sqrt{\frac{\pi}{2}}\,\frac{T_{Sys}}{\sqrt{\Delta\nu_{\mathrm{HF}}\,\tau_{\mathrm{LF}}}} \quad \ldots (3)$$

The sensitivity reduction is round about 10%. If a sinusoidal modulation is used, the sensitivity gets further reduced by a factor of 4/π.

3. Gain Modulated Dicke Receiver

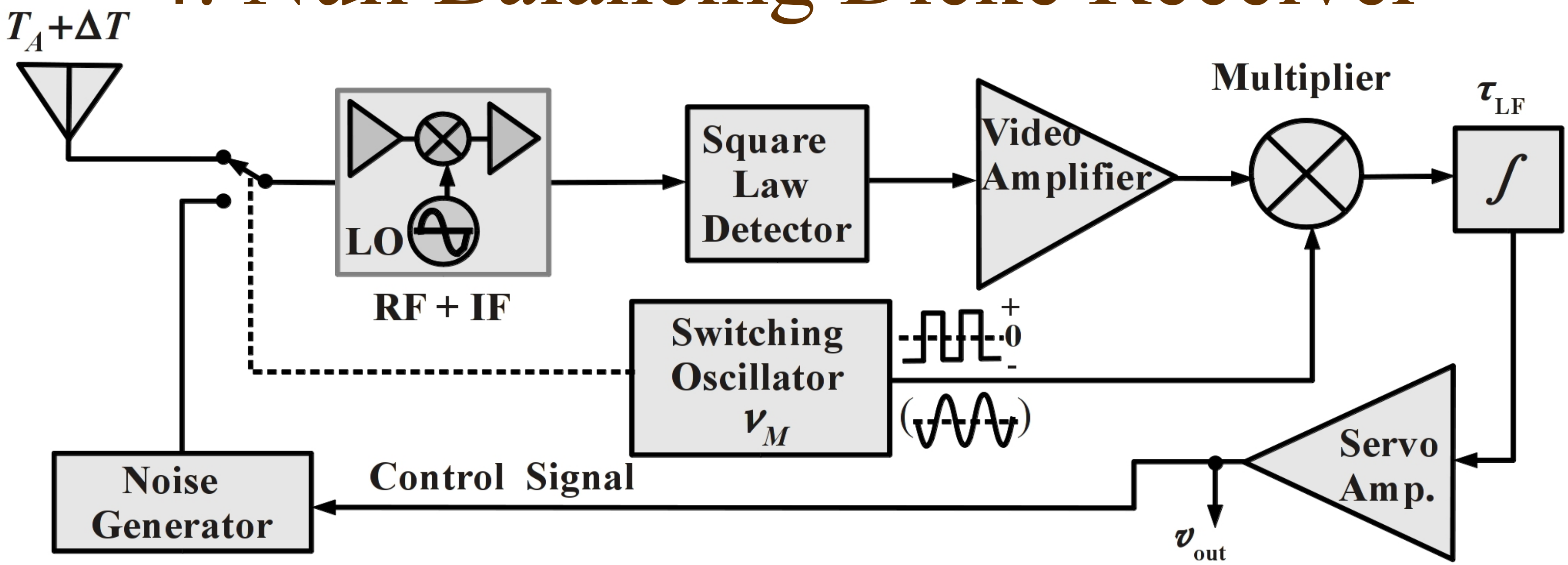


Two passive attenuators are used to control the gains of the paths from the (*i*) antenna, and from the (*ii*) noise generator to the square law detector. One of the attenuators, preferably in the path of the resistive noise source is variable. The attenuators are selected alternately by switching in synchronization with the input to the RF stage together with the modulating signal to the multiplier. The balance condition between the cold sky and the noise source is:

$$G_A\,(T_A + T_R) = G_C\,(T_C + T_R)$$

Here, T_A, T_R and T_C are respectively the antenna temperature (cold sky), receiver temperature and noise resistor temperature, and G_A and G_C are respectively the effective gains of the signal paths from the antenna and the resistor to the square law detector.

4. Null Balancing Dicke Receiver

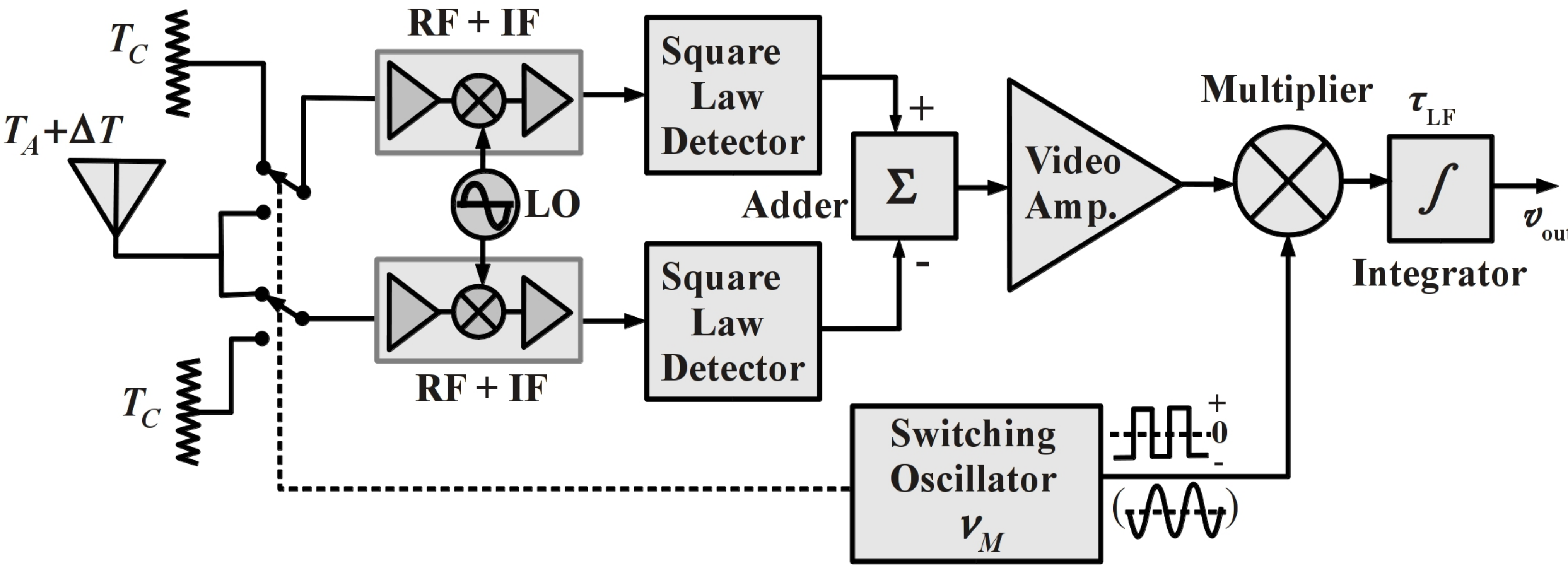


Dicke receivers described so far suffer from gain instability, when signals are present.

$$\Delta T_G = (T_A + \Delta T - T_C)\frac{\Delta G}{G_{\mathrm{HF}}} = \Delta T\,\frac{\Delta G}{G_{\mathrm{HF}}} \quad \text{since,} \quad T_A = T_C$$

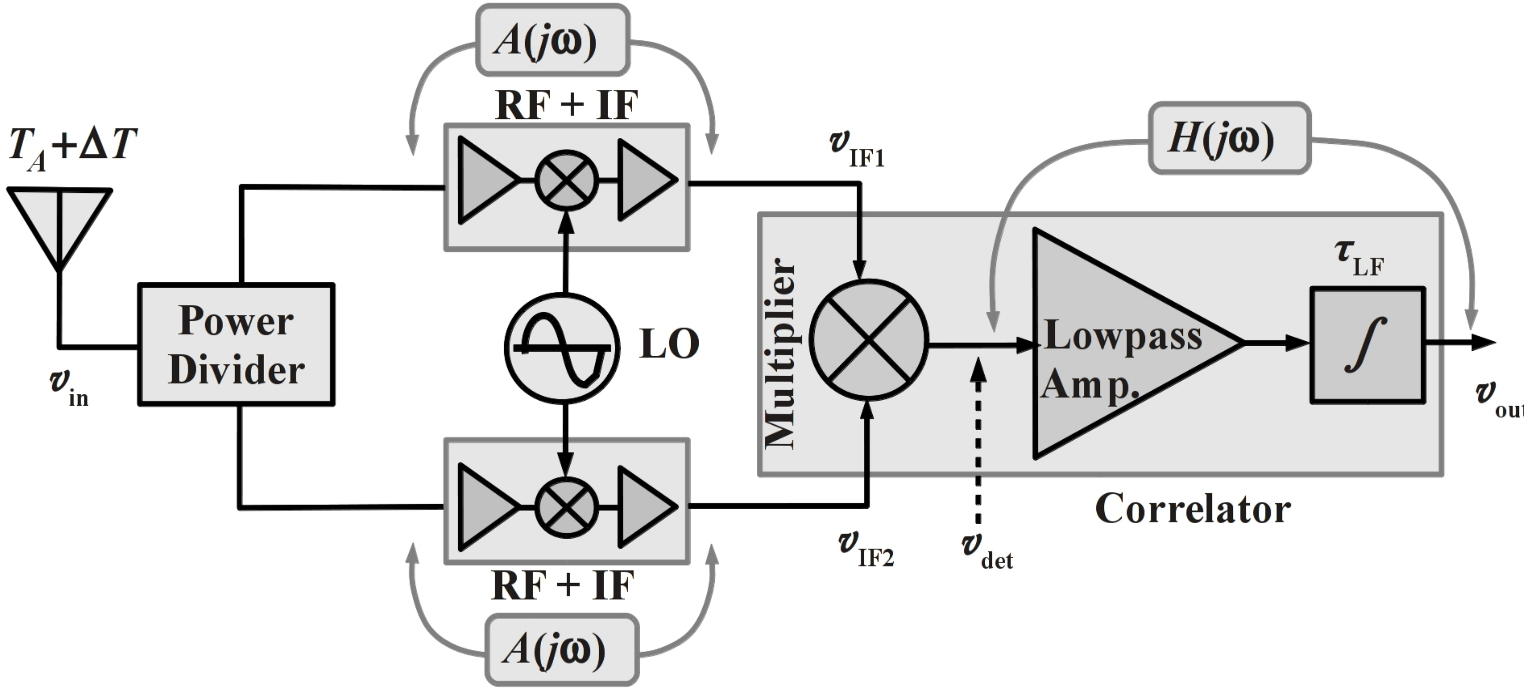
Further modifications were made by Machin, Ryle and Vonberg to the Dicke receiver which achieves the balanced condition for all times. It makes use of an adjustable noise source through a servo loop control mechanism. The power from this adjustable noise source is feedback controlled by the integrator output such that it is always zero. The control signal voltage v_{out} to the noise generator is proportional to the incremental temperature ΔT which can be recorded. The noise generator may be constructed using a noise diode.

5. Graham's Receiver



The signal power available in a simple Dicke receiver is only for half of the observation time. Hence, a large amount of signal goes unused. Graham's receiver avoids this loss. As shown, two receivers are used here which are time multiplexed to process the incoming signals. Hence, full observing efficiency is obtained. If the two receivers are alike, the sensitivity increases by a factor of $\sqrt{2}$.

6. Correlation Receiver-I



Antenna signal is split equally into two and fed to two identical heterodyne receivers (RF + IF) using a common LO. Hence, signal parts in these two receiver outputs are in phase. However, noise generated from the two receivers are uncorrelated. When multiplied with each other, the product of the actual signal parts are boosted while the noise parts are attenuated, thereby enhancing SNR. The low-pass filter cum amplifier boosts the lower frequencies and feeds them to the integrator. The SNR further improves after integration.

The two RF plus IF sections are identical having same receiver temperatures T_R.
System temperature T_{Sys} of each RF plus IF sections is $T_{Sys} = \frac{1}{2}T_A + T_R \quad \ldots (1)$

6. Correlation Receiver-II

The expected value $E[V_{out}]$ of the product v_{det} from v_{IF1} and v_{IF2} resulting from a Gaussian input v_{in} can be shown as:

$$E\left[V_{\text{out}}\right] = \frac{1}{4}k\left(T_A + \Delta T\right)\int_{-\infty}^{\infty}\left|A\left(j\omega\right)\right|^2 d\nu \qquad \ldots (2)$$

Here $A(j\omega)$ represents the frequency response of each of the RF plus IF sections, T_A is antenna temperature of cold sky, ΔT is the incremental temperature when moved from cold sky to a source, k is Boltzmann constant and $\omega = 2\pi\nu$.

Let σ_{out} be the standard deviation of v_{out}. If $\Delta T << T_A$, the relative deviation σ^*_{out} can be expressed as:

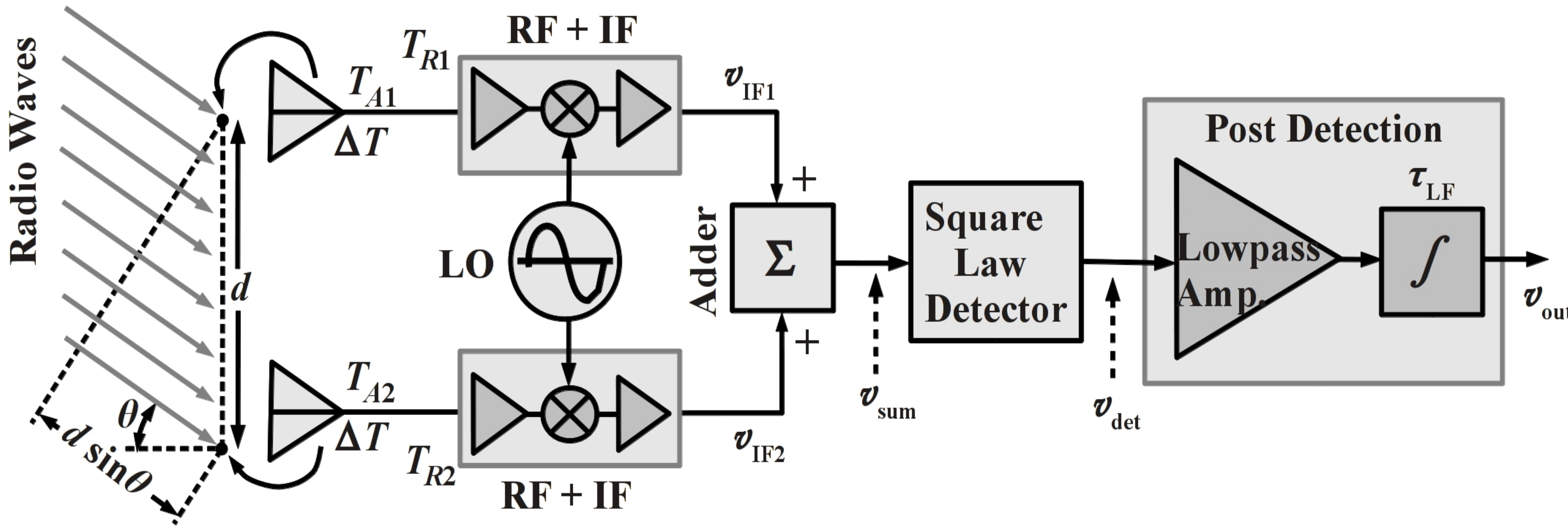
$$\sigma^*_{\text{out}} = \frac{\sigma_{\text{out}}}{E\left[V_{\text{out}}\right]} = \frac{2T_{Sys}}{\Delta T}\sqrt{\frac{\Delta\nu_{\text{LF}}}{\Delta\nu_{\text{HF}}}}\sqrt{1 + \left(\frac{T_A}{2T_{Sys}}\right)^2} \qquad \ldots (3)$$

where $\Delta\nu_{LF}$ is the band-width of the lowpass amplifier and $\Delta\nu_{HF}$ is the band-width of RF plus IF section. The sensitivity ΔT_{min} is given as:

$$\Delta T_{\text{min}} = \sqrt{2}\frac{T_{Sys}}{\sqrt{\Delta\nu_{\text{HF}}\ \tau_{\text{LF}}}}\sqrt{1 + \left(\frac{T_A}{2T_{Sys}}\right)^2} \qquad \ldots (4)$$

where τ_{LF} is the effective integration time.

7. Additive Interferometer Receiver-I



We have already seen the basic additive radio interferometer. Here we discuss its receiver also known as simple interferometer receiver. A pair of identical antennas connected to two identical pre-adder stages (RF + IF) having a common LO. The outputs are summed and sent to a square law detector. This is followed by post detection (low-pass amplifier and integrator).

The summed output voltage is $v_{\mathrm{sum}} = v_{\mathrm{IF1}} + v_{\mathrm{IF2}}$ … (1)

A phase difference may exist between the signals arriving from the two antennas depending on the source direction, but the signal magnitudes will be identical. v_{sum} is maximum if we satisfy: $d \sin\theta = n\lambda$ … (2)

v_{sum} is minimum if we satisfy: $d \sin\theta = \frac{(2n-1)\lambda}{2}$ … (3)

7. Additive Interferometer Receiver-II

For a discrete radio source, the signal (incremental) and noise (cold sky) powers from antenna-1 are respectively $k\,\Delta T\,\Delta\nu_{\mathrm{HF}}$ and $k\,T_{A1}\Delta\nu_{\mathrm{HF}}$, where k is the Boltzmann constant and $\Delta\nu_{\mathrm{HF}}$ is the available pre-adder bandwidths. For antenna-2, these are $k\,\Delta T\,\Delta\nu_{\mathrm{HF}}$ and $k\,T_{A2}\,\Delta\nu_{\mathrm{HF}}$. The receiver noise powers are $k\,T_{R1}\Delta\nu_{\mathrm{HF}}$ and $k\,T_{R2}\Delta\nu_{\mathrm{HF}}$. The noise power quantities $k\,T_{A1}\Delta\nu_{\mathrm{HF}}$, $k\,T_{A2}\Delta\nu_{\mathrm{HF}}$, $k\,T_{R1}\Delta\nu_{\mathrm{HF}}$ and $k\,T_{R2}\Delta\nu_{\mathrm{HF}}$ are assumed to be independent. Hence, in absence of a discrete radio source, the square law detector receives a noise power W_{det} given as:

$$W_{\mathrm{det}} = k\,G_{\mathrm{HF}}\,(T_{Sys1} + T_{Sys2})\,\Delta\nu_{\mathrm{HF}} \quad \ldots (4)$$

where, G_{HF} is the gain of the pre-adder stages, T_{Sys1} and T_{Sys2} are respectively the system temperatures of the two pre-adder sections given as:

$$T_{Sys1} = T_{A1} + T_{R1} \quad \ldots (5)$$

$$T_{Sys2} = T_{A2} + T_{R2} \quad \ldots (6)$$

Fluctuating noise power output W_{LF} from integrator is:

$$W_{\mathrm{LF}} = G_{\mathrm{LF}}\,2C'k^2\,(T_{Sys1} + T_{Sys2})^2\,\Delta\nu_{\mathrm{HF}}\,\Delta\nu_{\mathrm{LF}} \quad \ldots (7)$$

where, $\Delta\nu_{\mathrm{LF}}$ is the post detection band-width and C' is a constant.

7. Additive Interferometer Receiver-III

For a weak discrete source, the noise voltages from each IF amplifier are proportional to $\sqrt{(k\ \Delta T \Delta v_{HF})}$. If these two are in phase, the integrator output voltage v_{out} is proportional to $4k\ \Delta T \Delta v_{HF}$. If there exists a phase difference φ between the two, then v_{out} becomes proportional to $4k\ \Delta T \Delta v_{HF}(1+\cos\varphi)$. The integrator output power is: $W_{\text{out}} = G_{\text{LF}}\, C'\ \left[2k\, \Delta T\ \Delta \nu_{\text{HF}}\, (1 + \cos\phi)\right]^2 \quad \ldots (8)$

The sensitivity ΔT_{min} of the additive interferometer receiver is:

$$\Delta T_{\text{min}} = \frac{T_{Sys1} + T_{Sys2}}{2\,(1 + \cos\phi)\,\sqrt{\Delta\nu_{\text{HF}}\,\tau_{LF}}} \qquad \ldots (9)$$

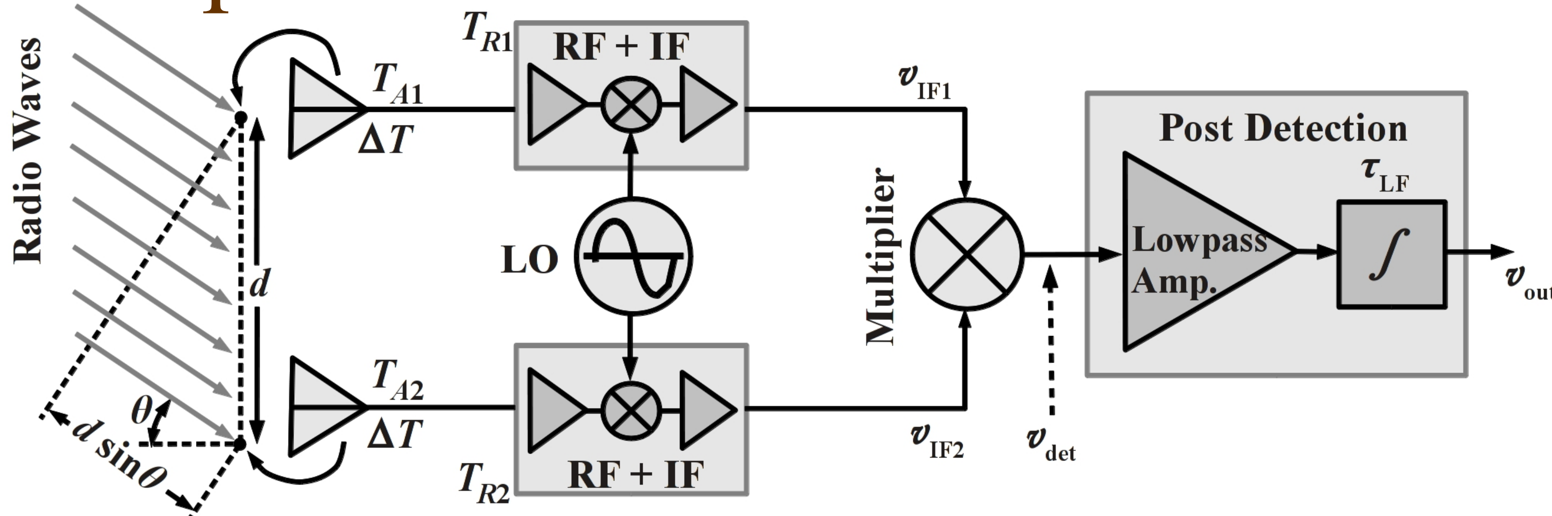
where, $t_{LF} = 1/(2\Delta v_{LF})$ is the effective integration time.

If $T_{Sys1} = T_{Sys2} = T_{Sys}$, we obtain a simplified expression as:

$$\Delta T_{\text{min}} = \frac{T_{Sys}}{(1 + \cos\phi)\,\sqrt{\Delta\nu_{\text{HF}}\,\tau_{LF}}} \qquad \ldots (10)$$

The simple additive interferometer have similar drawbacks of a total power receiver.

8. Multiplicative Interferometer Receiver-I



IF output voltages v_{IF1} and v_{IF2} are multiplied. Since, noise from pre-multiplier stages (RF + IF) are uncorrelated, they contribute very little to output voltage v_{out}. But, source signal contributions are correlated and contribute to v_{out}.

For a weak discrete source, noise voltages from each IF amplifier are proportional to $(k\,\Delta T\,\Delta\nu_{HF})^{0.5}$, where $\Delta\nu_{HF}$ is pre-multiplier band-width.

$$v_{IF1} = K_1 \sqrt{k\,\Delta T\,\Delta\nu_{HF}} \quad \ldots (1) \qquad v_{IF2} = K_1 \sqrt{k\,\Delta T\,\Delta\nu_{HF}} \quad \ldots (2)$$

Here, K_1 is the constant of proportionality (assumed identical for both paths).

Thus the detector output voltage v_{det} (from multiplier detector) is proportional to $k\,\Delta T\,\Delta\nu_{HF}\cos\varphi$, where φ is phase difference between v_{IF1} and v_{IF2}.

$$v_{det} = v_{IF1}\, v_{IF2} \cos\phi = K_2\,(k\,\Delta T\,\Delta\nu_{HF})\cos\phi \quad \ldots (3)$$

Here, K_2 is constant of proportionality for multiplier including K_1^2.

8. Multiplicative Interferometer Receiver-II

Signal power output from the integrator (delta function at zero frequency):

$$W_{\text{out}} = G_{LF}\, C' \; [k\, \Delta T\, \Delta\nu_{\text{HF}}]^2 \cos^2\phi \quad \ldots (4)$$

Noise power output from integrator (distributed noise across $\Delta\nu_{\text{LF}}$):

$$W_{\text{LF}} = G_{\text{LF}}\, 2C'k^2\, T_{Sys1}\, T_{Sys2}\, \Delta\nu_{\text{HF}}\, \Delta\nu_{\text{LF}} \quad \ldots (5)$$

Here, $\Delta\nu_{\text{LF}}$ is the post detection band-width and C' is a constant.

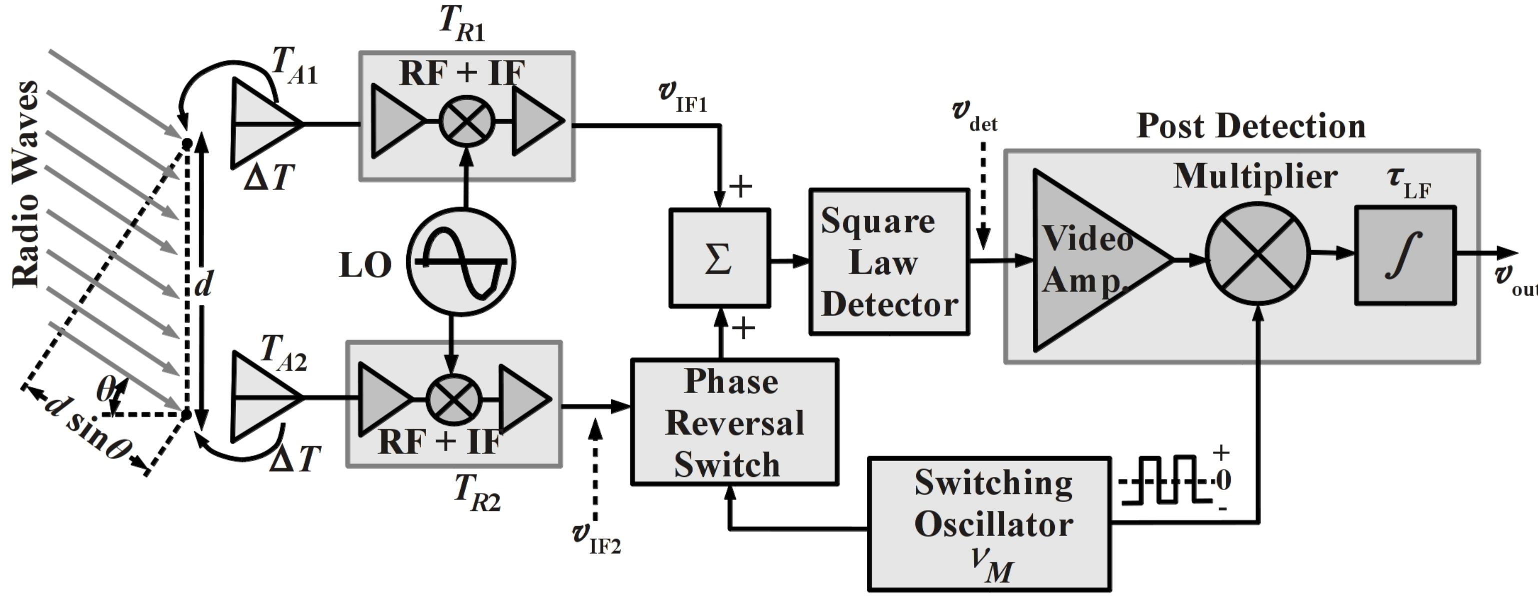
Sensitivity is: $$\Delta T_{min} = \frac{1}{\cos\phi}\sqrt{\frac{T_{Sys1}\, T_{Sys2}\, \Delta\nu_{\text{LF}}}{\Delta\nu_{\text{HF}}}} \quad \ldots (6)$$

If $T_{Sys1} = T_{Sys2} = T_{Sys}$, we obtain a simplified expression as:

$$\Delta T_{min} = \frac{1}{\cos\phi}\,\frac{T_{Sys}}{\sqrt{2\, \Delta\nu_{\text{HF}}\, \tau_{\text{LF}}}} \quad \ldots (7)$$

Sensitivity is $2\sqrt{2}$ times that of Dicke receiver. Since correlated noise alone can contribute to v_{out}, the gain instability seldom affects the sensitivity. However, gain variations can affect the receiver calibrations. Random phase variations in the pre-multiplier stage can also affect v_{out}. Scintillations in ionosphere also produces phase variations which is similar to the phase variation problem and can reduce the sensitivity in similar fashion. Since RF switch is not used at its inputs, extra loss of sensitivity due to increase in system temperature is avoided.

9. Phase Switched Receiver



It also uses the correlation principle. Two identical heterodyne systems (RF + IF) fed from a common LO are connected to the two antennas of the interferometer. The IF signal v_{IF2} goes through a phase reversal switch driven at a frequency ν_M. If v_{IF1} and v_{IF2} are uncorrelated, the switching will have no effect on the square law detector output voltage v_{det}. When v_{IF1} and v_{IF2} contain some correlated components, v_{det} is different for $v_{IF1}+v_{IF2}$ than for $v_{IF1}-v_{IF2}$. This implies that v_{det} changes at a frequency ν_M because of the correlated signal. If we assume that the desired signals alone are correlated, then the sensitivity of the phase switched receiver becomes same as that of Dicke receiver provided the latter also use a similar low frequency section.

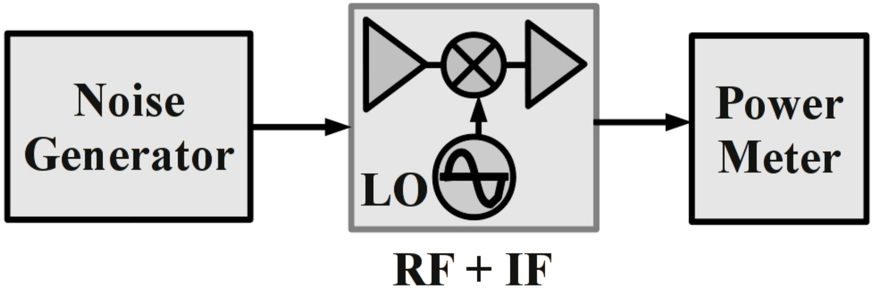
Comparisons of Sensitivities of Radiometers

$$\Delta T_{\min} = K_s \frac{T_{Sys}}{\sqrt{\Delta\nu_{\mathrm{HF}}\,\tau_{\mathrm{LF}}}}$$

S.N.	Receiver Type	Sensistivity K_s
1.	Total power receiver.	1
2.	Dicke receiver using square wave modulation, broad-band video amplifier followed by square wave multiplication.	2
3.	Dicke receiver using using square wave modulation, narrow-band video amplifier followed by sine wave multiplication.	$\pi/\sqrt{2} \approx 2.22$
4.	Dicke receiver using using sine wave modulation, narrow-band video amplifier followed by sine wave multiplication.	$2\sqrt{2} \approx 2.83$
5.	Graham's receiver.	$\sqrt{2} \approx 1.414$
6.	Correlation receiver having a small antenna noise in comparison to the receiver noise.	$\sqrt{2} \approx 1.414$
7.	Additive interferometer with identical antennas.	$1/2 = 0.5$
8.	Multiplicative interferometer with identical antennas.	$1/\sqrt{2} \approx 0.71$
9.	Phase switched interferometer with identical antennas, square wave switching and square wave multiplication.	2

Calibration of Receiver Temperature

Receiver noise temperature can be measured by connecting its input to a noise generator and measuring the output using a power meter.



Note: Internal impedance of noise generator must be same as of the antenna and should not fluctuate with time.

T_0 – Physical temperature of the noise generator.

T_G – Tempereature out of the noise generator.

Unfired condition: Generator temp. $T_G = T_0$ and power meter reading is P_{off}.

Fired condition: Generator temp. is T_G and power meter reading is P_{on}.

The ratio of these two readings is: $$\frac{P_{\text{on}}}{P_{\text{off}}} = \frac{T_G + T_R}{T_0 + T_R} \qquad \ldots (1)$$

Receiver temperature is: $$T_R = \frac{P_{\text{off}} T_G - P_{\text{on}} T_0}{P_{\text{on}} - P_{\text{off}}} \qquad \ldots (2)$$

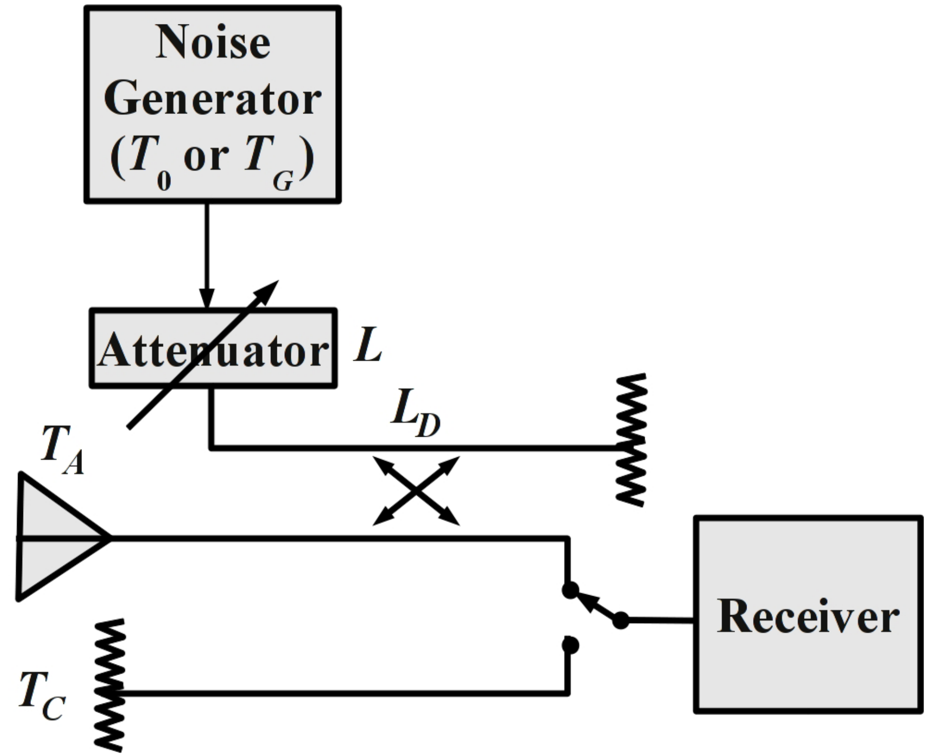
If the noise generator is adjustable, T_G can be adjusted so that $P_{\text{on}} = 2P_{\text{off}}$ and T_R may be expressed as: $$T_R = T_G - 2T_0 \qquad \ldots (3)$$

Calibration of Radiometers -I

Standard noise sources are used with receiving systems for calibrations. Types:

- (*i*) Internal calibration.
- (*ii*) External calibration.

Case (*i*) A noise generator is connected permanently to the receiving system using a directional coupler.



Unfired condition:

Generator temp. $T_G = T_0$

Fired condition:

Generator temp. is T_G

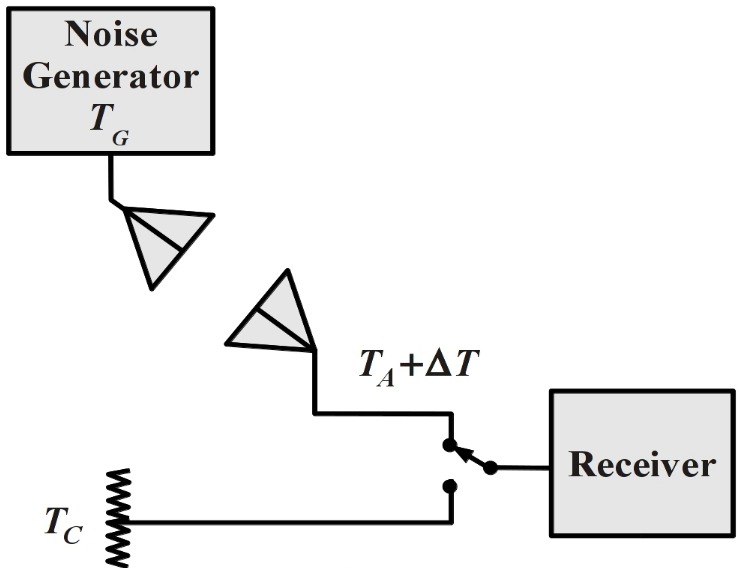
Generally, T_0 is close to the ambient temperature (300 K). A variable attenuator is used to control the noise flow to the receiver. If the attenuation is L and the coupling loss of the directional coupler is L_D, the excess noise temperature ΔT_E added to system can be estimated as:

$$\Delta T_{\mathrm{E}} = (T_G - T_0)\, G_{\mathrm{attn}}\, G_{\mathrm{dc}} = (T_G - T_0)\left(\frac{1}{L}\right)\left(\frac{1}{L_D}\right) = \frac{T_G - T_0}{L\, L_D} \quad \dots (1)$$

G_{attn} & G_{dc} - Out-to-in power ratios of attenuator and directional coupler.

Calibration of Radiometers -II

Case (*ii*) Noise generator is connected to an auxiliary antenna which radiates. It is received at a distance by the telescope antenna. Attenuation between telescope antenna and auxiliary antenna must be accurately known for absolute calibration. A radio source whose flux-density is known accurately can also be used for calibration, provided the effective aperture area A_e of the telescope antenna is accurately known.



The incremental temperature ΔT is given as:

$$\Delta T = \frac{A_e}{2k}\left(S - S_{\rm sky}\right) \quad \ldots (1)$$

Here, S is the spectral flux-density near the antenna aperture due to a point source and k is Boltzmann constant and $S_{\rm sky}$ is cold sky flux density. The source calibration temperature $T_{\rm cal}$ is given as:

$$T_{\rm cal} = \frac{S\, A_e}{2k} \quad \ldots (2)$$

Receiver calibrations are essential for providing an absolute scale of antenna temperature. Since gain and noise temperature of the receiver may vary during observation, calibration checks must be frequently made in between and also before and after observation.

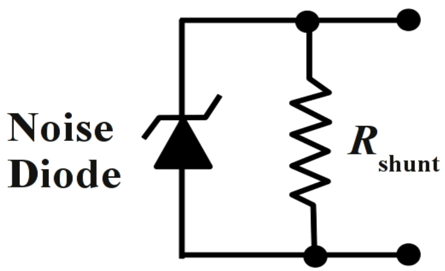
Noise Generators in Calibration-I

Several noise sources are available for calibration, but commonly used are:

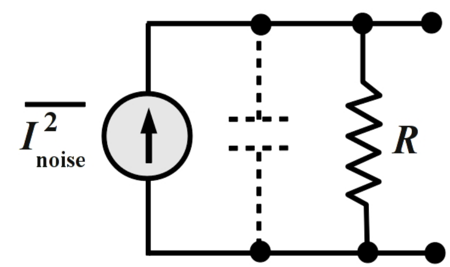
- (*i*) Thermally controlled resistors.
- (*ii*) Current controlled noise diodes.

Case (*i*) The noise power generated by the resistor is: $W_{\rm Res} = k\,T\,\Delta\nu$ … (1)
where k is Boltzmann constant, T is physical temperature of resistor and $\Delta\nu$ is band-width across which noise is measured. By changing resistor's temperature, equivalent amount of noise can be generated.

Case (*ii*) When large noise temperatures (1000K and above) are required, noise diodes are used. These are operated in the avalanche breakdown region and can work up-to 20 GHz. If diode biasing current is I_0, the mean square noise current is: $\overline{I^2_{\rm noise}} = 2\,I_0\,e^{-}\,\Delta\nu$ … (2) where, e^- is charge of an electron.

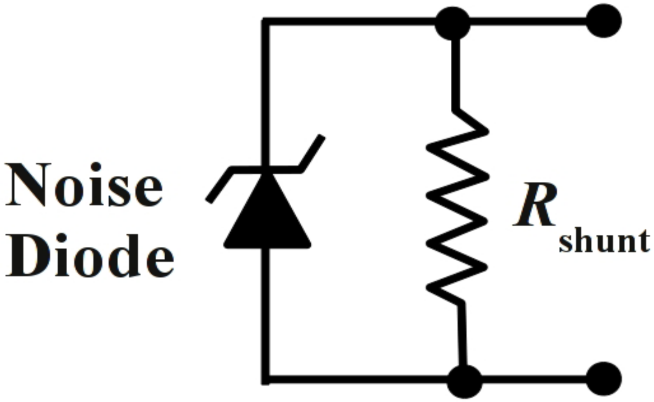


Equivalent RF circuit.

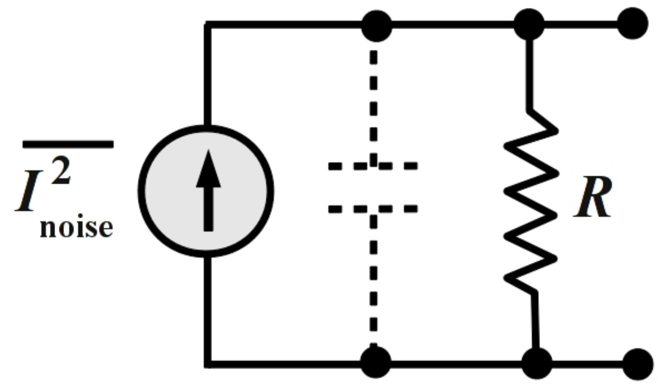


$R_{\rm shunt}$ is used for matching with transmission line. In RF equivalent circuit, R is the effective resistance. It is same as the characteristic impedance (resistive) of the transmission line. Junction capacitance of the diode (dotted lines) decides the maximum frequency of usage.

Noise Generators in Calibration-II



Equivalent RF circuit.



The output from the noise generator consist of two components:

(*i*) Noise due to diode biasing current I_0.

(*ii*) Noise due to physical temperature T_0 of the matching resistor.

The deliverable noise power W_G fom the noise generator to receiver under calibration is:

$$W_G = \left(\frac{\overline{I^2_{noise}}}{4} R \right) + (k\,T_0\,\Delta\nu) = \frac{I_0 e^{-} R\,\Delta\nu}{2} + k\,T_0\,\Delta\nu \qquad \ldots (3)$$

Here, $k\ T_0\ \Delta\nu$ is generated by the matching resistor due to its physical temperature T_0. The equivalent noise temperature T_G of the noise generator is:

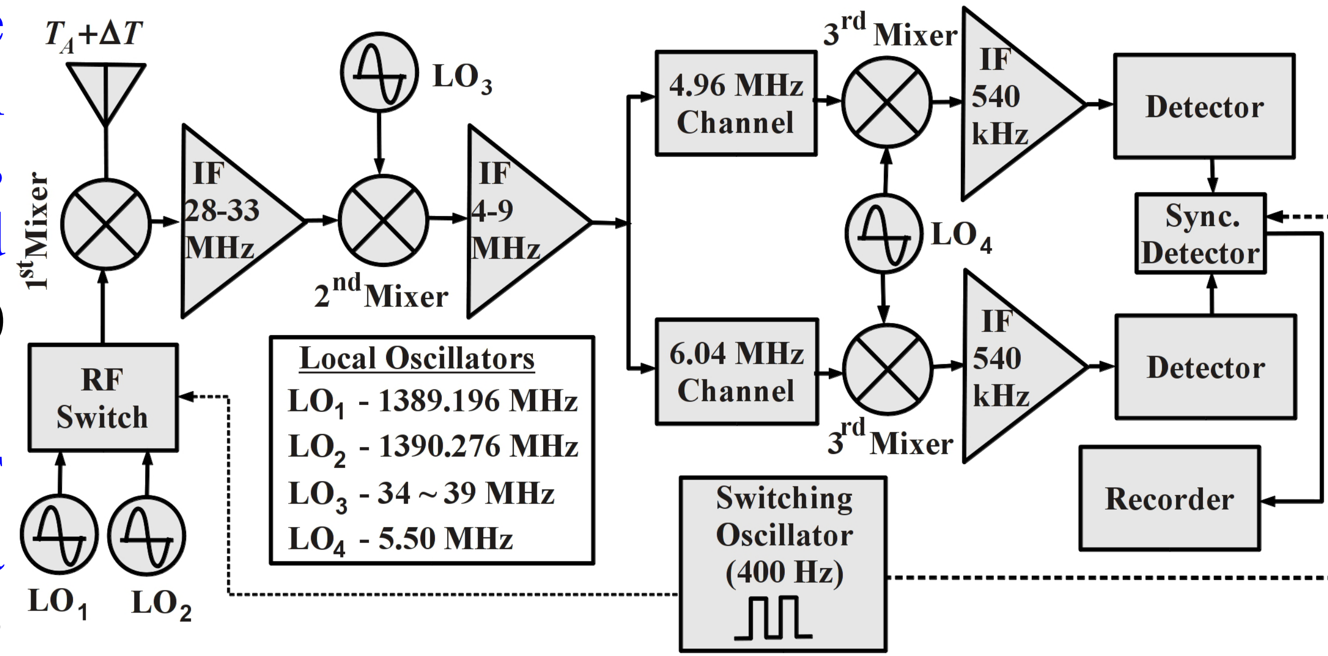
$$T_G = \frac{W_G}{k\Delta\nu} = \frac{I_0 e^{-} R}{2k} + T_0 \simeq (20 I_0 R + 1)\,T_0 \qquad \ldots (4)$$

Spectral Line Radiometers

Special radiometers were developed for correctly measuring the 1420 MHz line radiations from neutral hydrogen from Milky Way and other galaxies. Later, frequency resolution was increased by adding more number channels with reduced channel band-widths. This enabled to view the details of the spectra within a narrow band-width covering the spectral line.

Though modern radio telescopes use interferometers for spectral line observations, we describe one of the earlier versions known as ***Frequency-Switched Radiometer*** which might interest the students.

Frequency Switched Radiometer-I

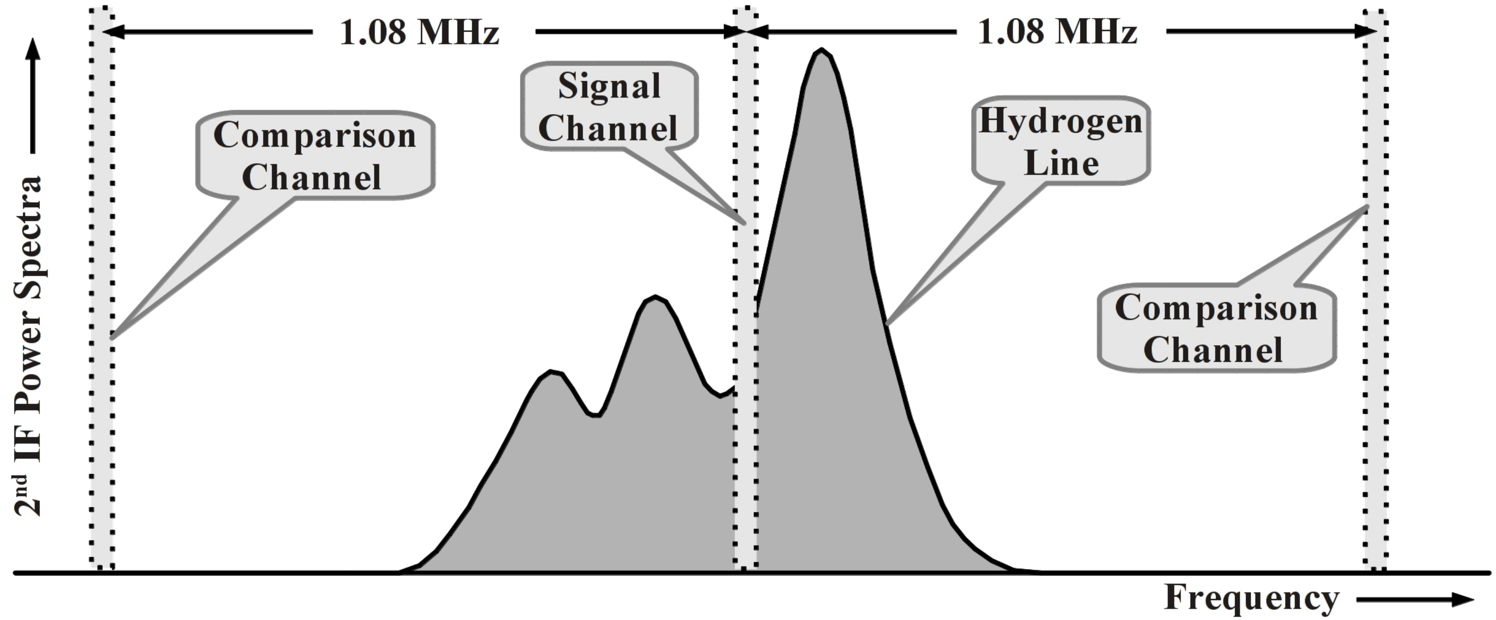


Antenna signals are alternatively mixed with two crystal local oscillators LO_1 (1389.196 MHz) and LO_2 (1390.276 MHz) at 400 Hz using a RF switch. Mixer output is fed to a IF filter amplifier having a frequency band of 28 to 33 MHz. After amplification it goes into the second mixer, whose local oscillator LO_3 (set at 35.764 MHz) can be swept between 34 to 39 MHz. The second IF output is amplified within a range of 4 to 9 MHz using another filter amplifier. It is then fed to two separate channel filters centered at 4.96 MHz and at 6.04 MHz. Difference between LO_1 and LO_2 is 1.08 MHz. Same difference is maintained between the channels. These two channel outputs are mixed with LO_4 (5.50 MHz) over a band-width of 35 kHz and processed through two IF filter amplifiers at 540 kHz. These are next sent to two detectors. The detected signals signals are sent to a synchronous detector which is also supplied with the 400 Hz switching signal.

Frequency Switched Radiometer-II

If a 1420 MHz signal is found, it will emerge through the 4.96 MHz channel when LO_1 operates, and through the 6.04 MHz channel when LO_2 operates.

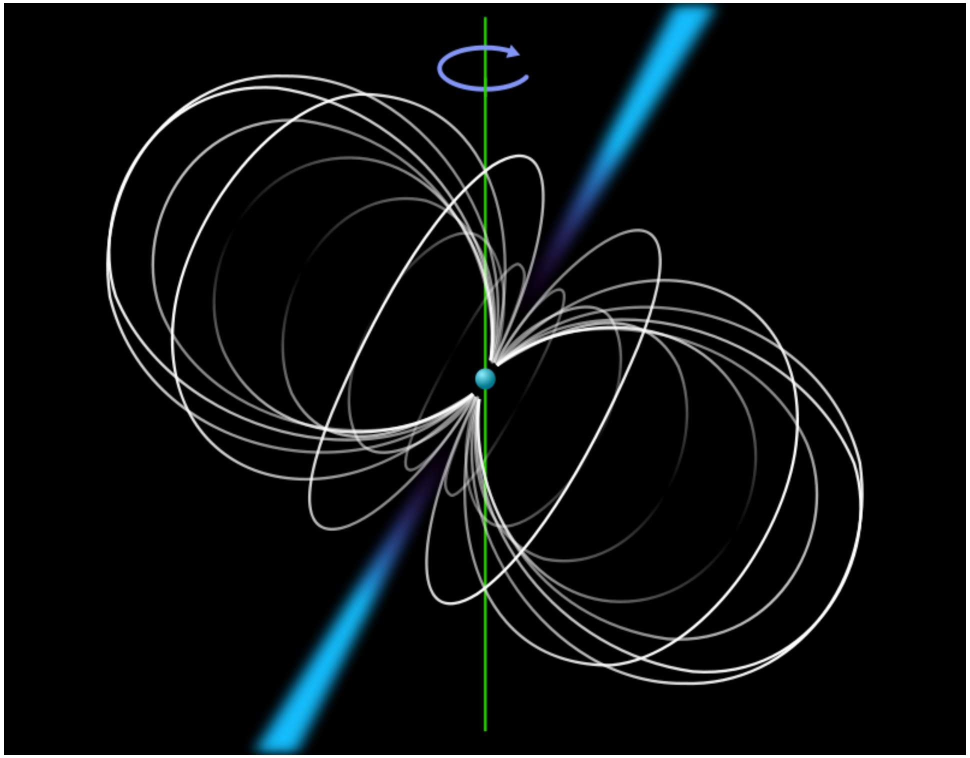
Location of channels with respect to a hydrogen line profile (spectra) at 2nd IF output.



LO_3 is tuned such that the information near 1420 MHz of the hydrogen line is captured by one of the two channels. The other channel lies 1.08 MHz below or above the first. The synchronous detector output measures the (*i*) difference between the noise power at 1420 MHz, and (*ii*) mean of the noise power at frequencies equally spaced below and above 1420 MHz. By suitably tuning LO_3, the sensitive band of the receiver may be swept across the hydrogen line profile. Switching can improve the receiver's overall stability since the effects of receiver noise are minimized. The line is observed in both switching positions to avoid loss of sensitivity.

Pulsars: Introduction-I

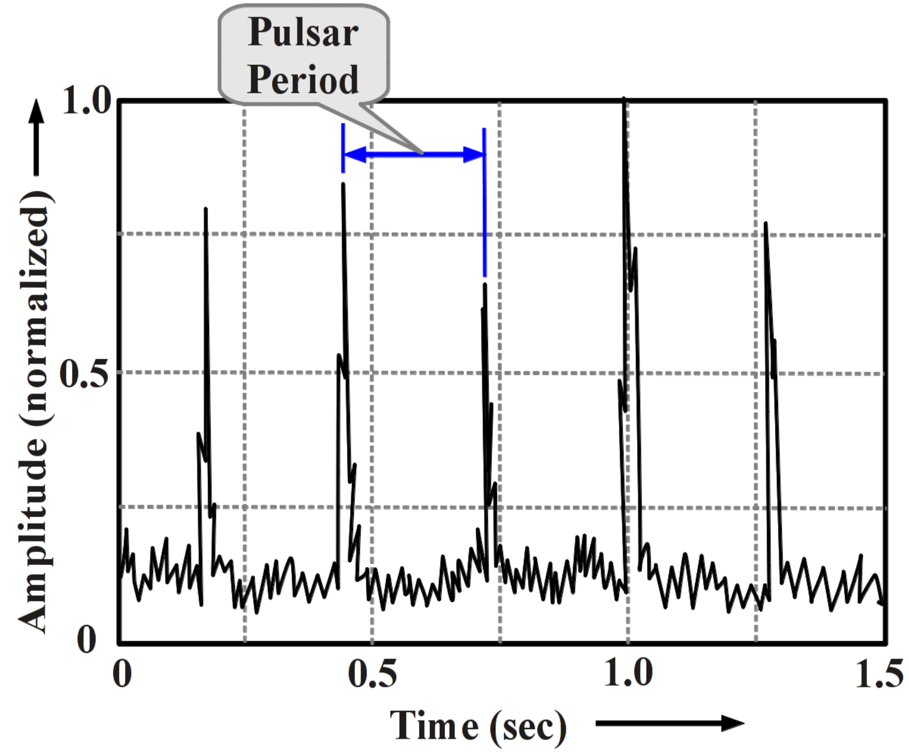
The core of a massive star after supernova collapses into a high speed rotating neutron star for it retains most of its angular momentum. A synchrotron electro-magnetic radiation beam is emitted along the magnetic axis, which spins along with its rotation. The magnetic axis of the pulsar determines the direction of the electromagnetic beam, which may not necessarily be the same as rotational axis. This misalignment causes the beam to be seen once for every rotation of the neutron star, leading to its ***pulsed*** nature.



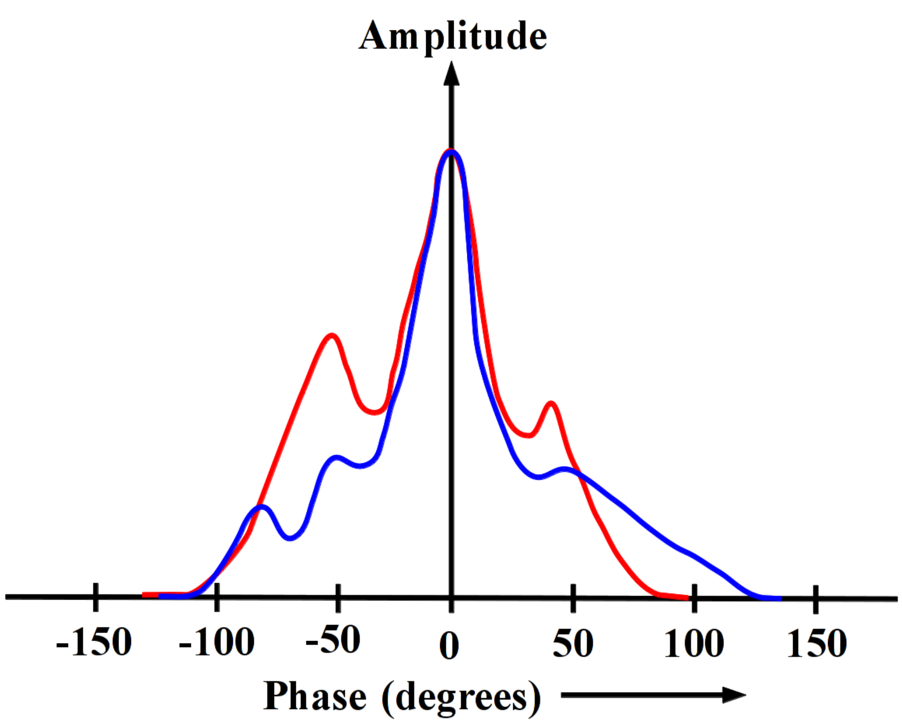
The rotational energy generates an electrical field from the movement of very strong magnetic field, resulting in the acceleration of protons and electrons on the star surface and the creation of an electromagnetic beam emanating from the magnetic poles. The rotation slows down with time as electromagnetic power is emitted. When its spin period slows down sufficiently, the radio pulsar mechanism is believed to turn off (10–100 million yrs).

Pulsars: Introduction-II

The time interval between two pulses is called as ***pulse period***. The pulses can look very different from each other in both shape and height as seen in the plot. There are two main types of pulsars. Those with periods of a few milliseconds are called the ***millisecond pulsars***. These pulsars' periods change slowly with time. The rest are known as ***ordinary pulsars***.



Pulses from a generic pulsar.



Generic averaged pulsar profiles at two frequency bands shown by blue and red curves.

Pulsar Receivers

Pulsars may be broadly grouped into two:

(*i*) Those whose pulses remain constant over many pulse periods.

(*ii*) Those whose pulses vary from pulse to pulse.

A common receiver hardware can be used for both. However, data handling of these can be completely different. To obtain integrated pulse characteristics, the receiver output must be averaged synchronously with the pulsar period and then record them at relatively large time intervals of the order of several minutes or more. This may be done on a computer or using a special hardware, provided the pulse period is accurately known before hand.

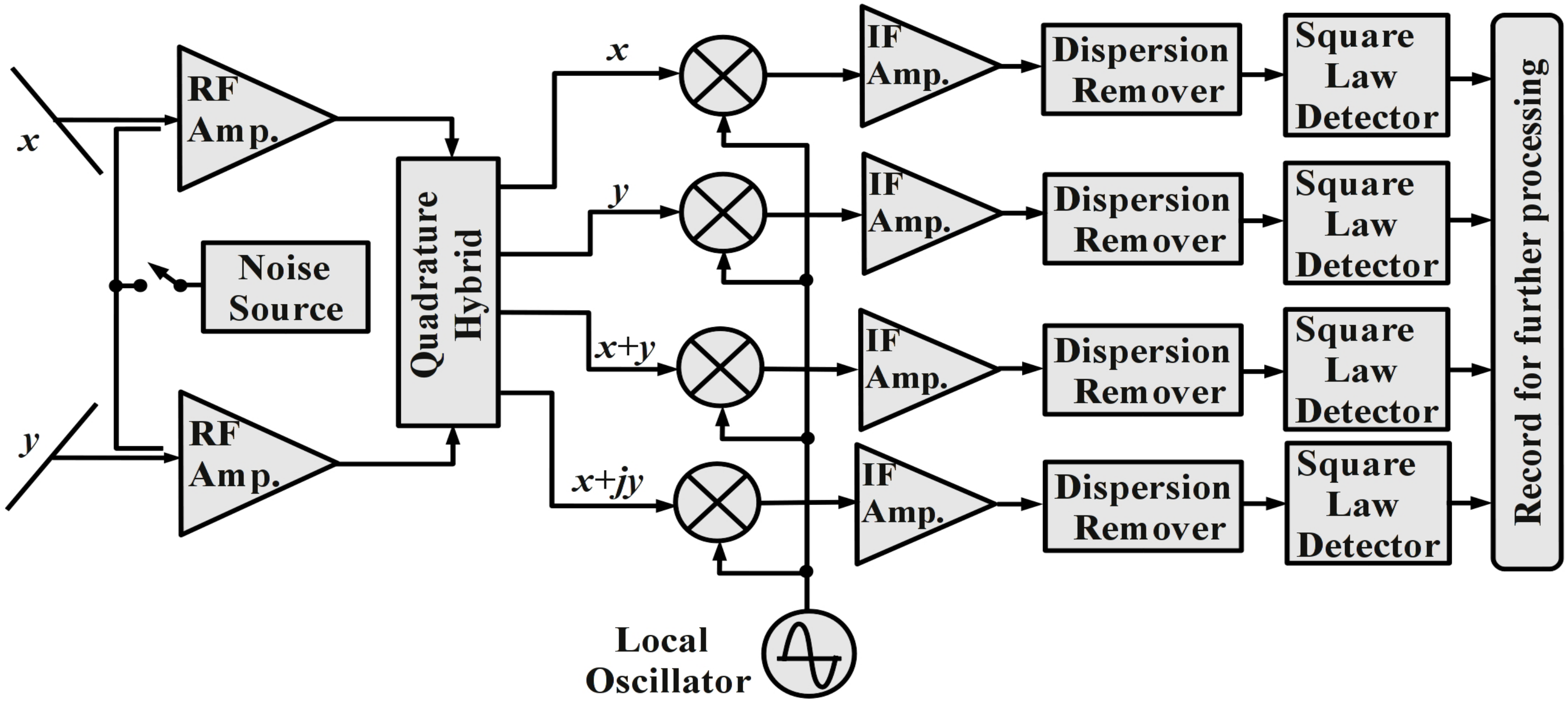
Pulsar receivers differ from conventional radio astronomy receivers in three ways:

(*i*) Synchronous detection (comparison switching) is avoided since the pulses themselves provide switching information.

(*ii*) Four identical signal channels are used to instantaneously determine the polarization states of the rapidly varying signals without feed rotation.

(*iii*) To compensate the dispersion effects from the inter-stellar medium, dispersion removers are used in all the four signal paths.

System Blocks of Pulsar Receivers



System blocks of a pulsar receiver.

Here x and y represent the two orthogonal linear polarizations of the antenna. The noise source is for simultaneous calibration of both polarizations. After amplification, the signals are added in a quadrature hybrid to generate $x+y$ and $x+jy$ components. All of these (x, y, $x+y$ and $x+jy$) are individually mixed with a common LO signal and respective IF are generated. These are amplified and passed through dispersion removers and then fed to square law detectors. The detector outputs are recorded (kept for further processing like pulse averaging etc.).

Pulsar: Dispersion Measure (DM)

The pulses get dispersed in the inter-stellar medium, which is frequency dependent. Hence use of large band-widths for increasing sensitivity is restricted. For a desired time resolution of $\Delta\tau$ (sec) the maximum usable band-width $\Delta\nu_{max}$ (MHz) is:

$$\Delta\nu_{max} = \Delta\tau \frac{d\nu}{dt} \approx 1.2 \times 10^{-4} \Delta\tau \, \nu^2 \, \mathrm{DM}^{-1} \quad \ldots (1)$$

Here, $d\nu/dt$ (MHz/sec) is the rate of change of frequency of the dispersed signal, ν is the observing frequency and DM (cm^{-3} pc) is the dispersion measure.

DM may be thought as representing the number of free electrons between us and the pulsar per unit area. Imagine a long tube of unit cross sectional area extending from us to the pulsar. DM is proportional to the number of free electrons inside this volume. In reality, DM actually represents the total column density of free electrons existing between the observer and the pulsar. It is the integral of the electron density along the line of site given as:

$$\mathrm{DM} = \int_0^D n_e(s) \, ds \qquad \ldots (2)$$

Here, D represents the distance between the pulsar and the observer, and n_e is the electron density of the ISM (inter stellar medium).

Pulsar: Dispersion Correction (De-Dispersion)

To remove the dispersion effects, a complex Fourier analysis of the IF signal is done. This complex spectrum is multiplied with a frequency dependent phase factor which is pre-computed from already known DM of the pulsar. Inverse Fourier transform of this gives the desired signal (free from the dispersion effects).

The dispersion correction methods may be broadly categorized into two:

(*i*) Incoherent de-dispersion.

(*ii*) Coherent de-dispersion.

Pulsar: (*i*) In-Coherent De-Dispersion-I

This is like separating the frequency channels of the incoming signal using a number of narrow-band filters and then applying different time delays based on their respective channel frequencies and finally combining them as output. These delays compensate for the frequency dependent delays applied by the inter-stellar medium to the signal. Hence the corrected signal channels are in phase which are added to obtain the actual signal.

The difference in arrival times Δt (ms) between a pulse received by a channel of frequency ν_{ch} (MHz) relative to the center frequency ν_0 of the of the observed band is approximately:

$$\Delta t = 4.15 \times 10^6 \left(\frac{1}{\nu_0^2} - \frac{1}{\nu_{\rm ch}^2} \right) \times {\rm DM} \quad \ldots (1)$$

where, DM is in $\mathrm{cm^{-3}\,pc}$.

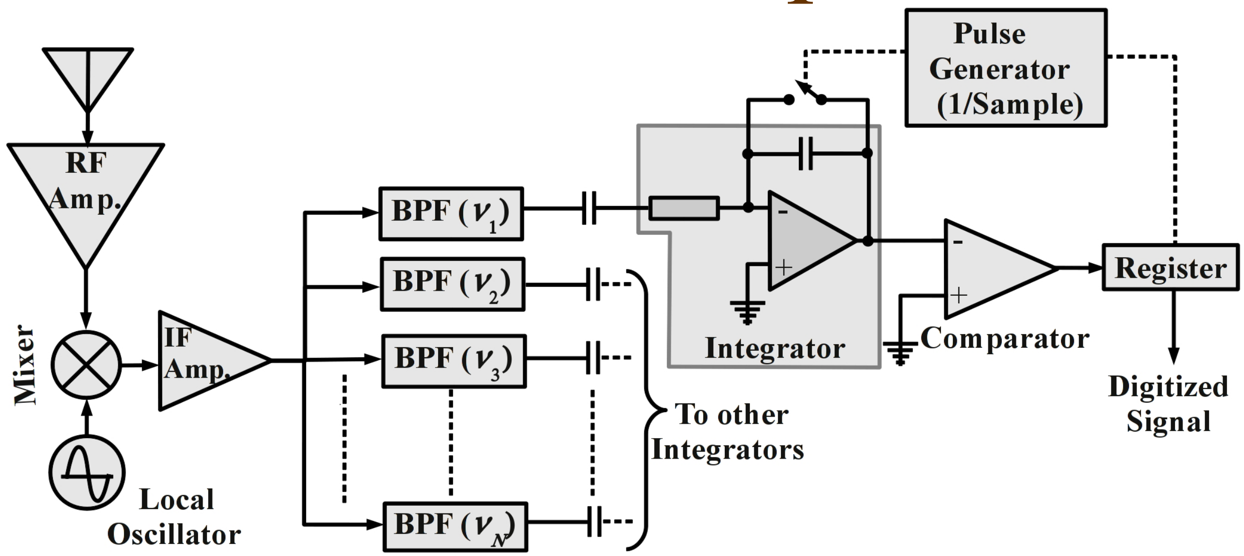
For a finite band-width near a center frequency ν_0 (MHz), the dispersive delay τ_{DM} (ms) across a frequency channel of band-width $\Delta\nu$ (MHz) with condition that $\nu >> \Delta\nu$ is:

$$\tau_{\rm DM} = 8.3 \times 10^6 \times {\rm DM} \times \Delta\nu \times \nu^{-3} \quad \ldots (2)$$

Hence, channel band-widths must be carefully chosen to prevent τ_{DM} to become a significant fraction of the pulse period.

Pulsar: (*i*) In-Coherent De-Dispersion-II

An ***analog filter-bank spectrometer*** using single bit digitization scheme. It is the most popular and simplest data acquisition device for incoherent de-dispersion.



Broad-band IF is split into N adjacent channels having center frequencies $\nu_1, \nu_2, \ldots, \nu_N$ followed by integration and digitization using a 1 bit sampler. Integrator output is the running mean of the signal. It is compared with 0 voltage inside the comparator. Its output is digital 0 or 1. Sampling is achieved by reading the comparator output in a register at a rate of $1/\tau_s$, where τ_s is the sampling period.

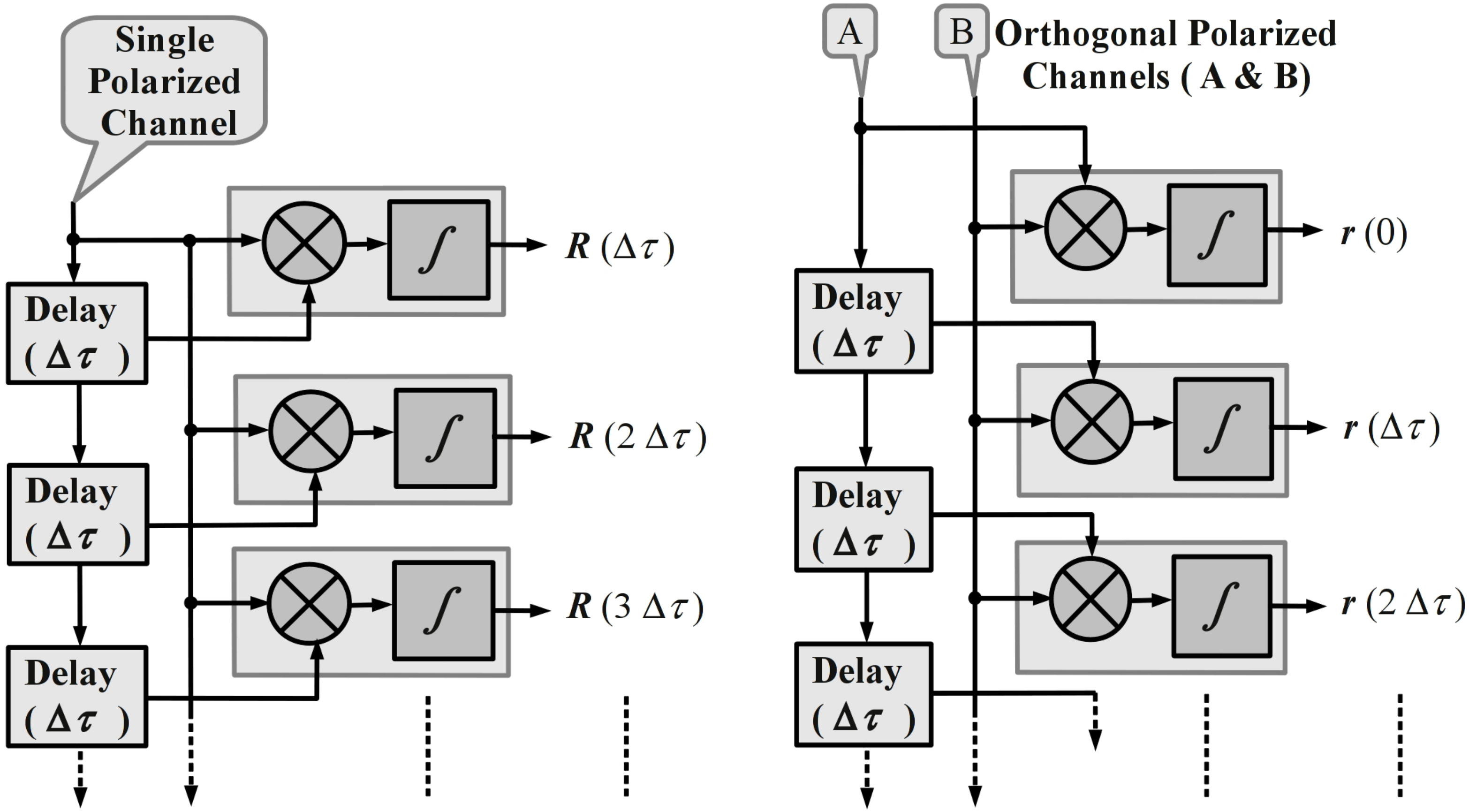
Pulses are also sent to integrator switch for discharging the capacitor in every cycle. The integrator acts as a LPF and rejects nearly all frequencies above $1/\tau_s$.

No anti-aliasing filters are used. All the frequency channels are digitized separately. The loss is about 20% from ideal sampling.

Very useful in large projects for pulsar search. Almost two thirds of today's know pulsars have been discovered using this technique.

Pulsar: (*i*) In-Coherent De-Dispersion-III

Functional blocks of a correlation spectrometer.



Auto-correlation.

Cross-correlation.

Any single polarized channel is split and multiplied with its delayed version followed by integration. The time delays used are integer multiples of $\Delta\tau$ which is the minimum time delay adopted. Cross-correlation products between the two orthogonal linear polarizations A and B are also generated as shown.

Pulsar: (i) In-Coherent De-Dispersion-IV

Let $R(\tau)$ represent the auto-correlation and $H(\nu)$ be its power spectrum. Let $r(\tau)$ be the cross-correlation. From Wiener-Khinchin theorem, it may be said that $R(\tau)$ obtained from an interferometer is the Fourier transform of the cosmic signal's power spectrum $H(\nu)$ having a band-width equal to the system pass-band. If H_{AA} and H_{BB} are respectively the power spectra for the channels A and B resulting from auto-correlations, then the total intensity $I = H_{AA}+H_{BB}$. Let H_{AB} and H_{BA} be the cross-correlation power spectra between A and B. Using these, the Stokes parameters I, Q, U and V from polarimetric observations using a dual polarized linear antenna-feed can be shown as:

$$\begin{bmatrix} I \\ Q \\ U \\ V \end{bmatrix} = \begin{bmatrix} H_{AA} + H_{BB} \\ H_{AA} - H_{BB} \\ 2\,\Re e\,[H_{AB}] \\ 2\,\Im m\,[H_{BA}] \end{bmatrix} \quad \ldots (1)$$

If the antenna-feed is dual circularly polarized, the Stokes parameters are:

$$\begin{bmatrix} I \\ Q \\ U \\ V \end{bmatrix} = \begin{bmatrix} H_{AA} + H_{BB} \\ 2\,\Im m\,[H_{BA}] \\ 2\,\Re e\,[H_{AB}] \\ H_{BB} - H_{AA} \end{bmatrix} \quad \ldots (2)$$

The Stokes parameters obtained using this method is simpler as compared to those obtained using analog filter banks. As a precaution, the auto and cross-correlations must be calibrated before analysis and interpretation.

Pulsar: (*ii*) Coherent De-Dispersion-I

Here, the phase information obtained from the telescope output voltage can be used to completely remove the phase change effects caused by ISM. The propagating signal through ISM is treated like a phase change caused by a phase filter having a transfer function $H(\nu)$.

Let a signal possess a band-width $\Delta\nu$ centered at a frequency ν_0. Let $v_{in}(t)$ and $v_{out}(t)$ be respectively the input and output voltages of of the phase filter. Let their Fourier transforms be $V_{in}(\nu)$ and $V_{out}(\nu)$ respectively. These quantities can be related together using $H(\nu)$ as:

$$V_{\text{out}}(\nu_0+\nu) = V_{\text{in}}(\nu_0+\nu)\, H(\nu_0+\nu) \quad \text{where, } |\nu| \le \Delta\nu/2 \qquad \ldots (1)$$

The objective is to regain the actual pulsar signal $v_{in}(t)$ from $v_{out}(t)$ with the ISM acting as the phase filter. The delay caused by ISM may be represented by phase rotations which depend on frequency and path length of travel. Hence, the phase change $\Delta\varphi$ can be shown as:

$$\Delta\phi = -k(\nu_0+\nu)\, d \qquad \ldots (2)$$

where $k(\nu)$ is the wave number at a frequency ν and d is the distance of travel. We now express $H(\nu)$ as:

$$H(\nu_0+\nu) = e^{-j\, k(\nu_0+\nu)\, d} \qquad \ldots (3)$$

Pulsar: (*ii*) Coherent De-Dispersion-II

The wave number in terms of the plasma frequency ν_p and cyclotron frequency ν_B is:

$$k(\nu_0+\nu) = \frac{2\pi}{c}(\nu_0+\nu)\sqrt{1-\frac{\nu_p^2}{(\nu_0+\nu)^2} \mp \frac{\nu_p^2\,\nu_B}{(\nu_0+\nu)^3}} \quad \ldots (4)$$

In ISM, $\nu_p \approx 2$ kHz and $\nu_B \approx 3$ Hz. For observations above 100 MHz, the last two terms of the above equation are of the orders of 10^{-10} and 10^{-18}. Hence, the wave number is approximately given as:

$$k(\nu_0+\nu) \approx \frac{2\pi}{c}(\nu_0+\nu)\left[1-\frac{\nu_p^2}{2(\nu_0+\nu)^2}\right] \quad \ldots (5)$$

Substituting (5) in (3) and after simplifications we obtain $H(\nu)$ as:

$$H(\nu_0+\nu) = e^{-j\frac{2\pi}{c}d\left[\left(\nu_0-\frac{\nu_p}{2\nu_0^2}\right)+\left(1+\frac{\nu_p^2}{\nu_0^2}\right)\nu-\frac{\nu_p^2}{2(\nu_0+\nu)^2}\nu^2\right]} \quad \ldots (6)$$

First term of exponent is destroyed by square law detector. Second term links to a time delay. It may be nullified by suitably shifting the arrival time. The final term is responsible for phase rotations in those bands which are quadratic in frequency. Importance of this term lies in recovering the original pulsar signal. Thus, (6) gets simplified to:

$$H(\nu_0+\nu) = e^{j\frac{2\pi}{c}d\left[\frac{\nu_p^2}{2(\nu_0+\nu)^2}\right]v^2} \quad \ldots (7)$$

Pulsar: (*ii*) Coherent De-Dispersion-III

A dispersion constant $\mathcal{D}$ is defined in relation to the plasma frequency ν_p. It is given as:

$$\mathcal{D} = \frac{\nu_p^2}{2\,c\,n_e} \qquad \ldots (8)$$

Here c is the speed of light and n_e is the average electron density along the line of site. The dispersion constant $\mathcal{D}$ may be related to the DM as:

$$\frac{\nu_p^2\,d}{2\,c} = \mathrm{DM} \times \mathcal{D} \qquad \ldots (9)$$

Relating (8) with (9), the transfer function H(ν) is finally expressed as:

$$H(\nu_0 + \nu) = \exp\left[\frac{j2\,\pi\,\mathcal{D}}{(\nu_0+\nu)^2\,\nu_0^2}\mathrm{DM}\,\nu^2\right] \qquad \ldots (10)$$

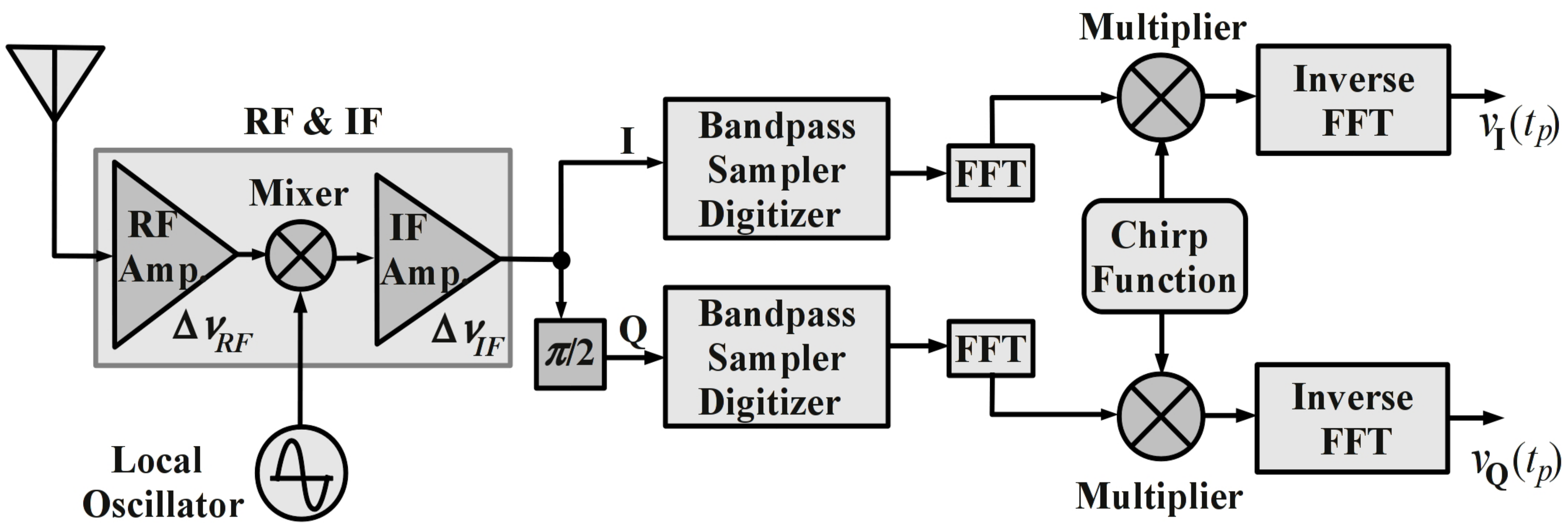
The inverse transform H^{-1} is now applied to the voltage data $V_{\text{out}}(\nu)$ to obtain originally emitted voltage $V_{\text{in}}(\nu)$ from the pulsar as:

$$V_{\text{in}}(\nu) = V_{\text{out}}(\nu)\,H^{-1}(\nu_0+\nu) \qquad \ldots (11)$$

This process however is slightly complicated for practical implementation since digitization is involved. Generally, H^{-1} is multiplied with a tapering function $T(\nu)$ which takes care of the practical problems. The resulting product $C(\nu_0,\nu)$ is known as ***chirp function***:

$$C(\nu_0,\nu) = T(\nu)\,H^{-1}(\nu_0+\nu) \qquad \ldots (12)$$

Pulsar: (*ii*) Coherent De-Dispersion-IV



Functional blocks of coherent de-dispersion receiver.

After down conversion, the IF output generates two orthogonal functions I and Q using a 90° phase shifter. These are band-pass sampled and digitized. Next FFT is performed over a fixed number of samples. These are then multiplied with a digitized Chirp function. Finally, inverse FFT is performed and the time domain digitized values of the signal are obtained, which are free from dispersion effects.

Pulsar: (*ii*) Coherent De-Dispersion-V

By combining the two orthogonal linear polarizations (X and Y) of the antenna-feed, the Stokes parameters I, Q, U and V are obtained as:

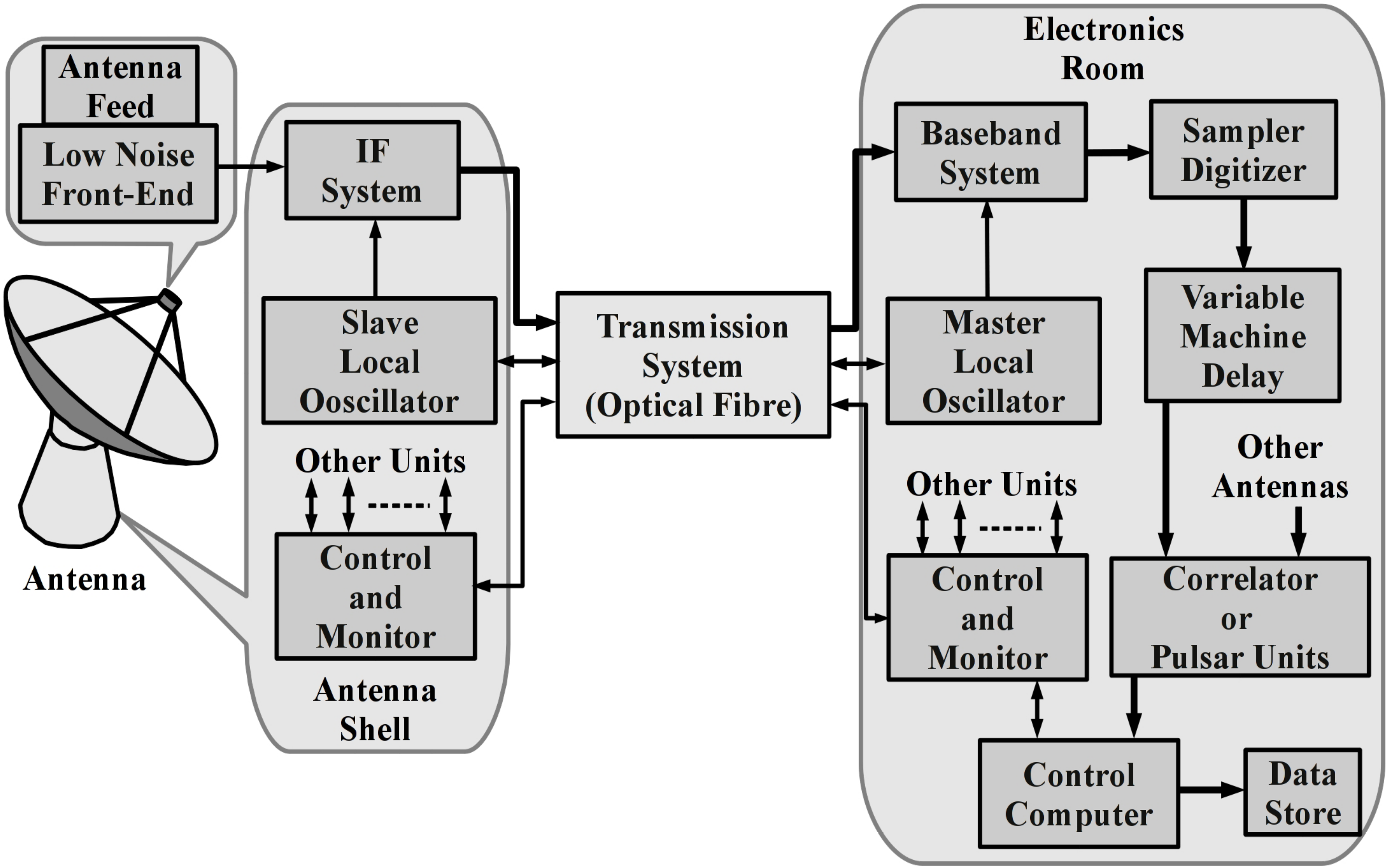
$$\begin{bmatrix} I \\ Q \\ U \\ V \end{bmatrix} = \begin{bmatrix} |X|^2 + |Y|^2 \\ |X|^2 - |Y|^2 \\ 2\,\mathrm{Re}\,[X^*\, Y] \\ 2\,\mathrm{Im}\,[X^*\, Y] \end{bmatrix} \qquad \ldots (13)$$

where, X^* represents the complex conjugate of X.

Similarly, for a dual circularly polarized antenna-feed with L and R as the left and right circularly polarized components, the Stokes parameters are obtained as:

$$\begin{bmatrix} I \\ Q \\ U \\ V \end{bmatrix} = \begin{bmatrix} |L|^2 + |R|^2 \\ 2\,\mathrm{Re}\,[L^*\, R] \\ 2\,\mathrm{Im}\,[L^*\, R] \\ |R|^2 + |L|^2 \end{bmatrix} \qquad \ldots (14)$$

A Generic Receiving Chain Layout-I



A generalized layout of the complete receiving and control chain from observatory building to antennas of a radio array.

A Generic Receiving Chain Layout-II

The low noise front-end system follows the antenna-feed. It contains RF amplifiers. Its output is brought through a RF cable inside the antenna shell where the remaining front-end electronics exist. IF signals are obtained after mixing with the LO signal. The LO is locked to a master oscillator, located inside observatory building. Other control units like servo for antenna azimuth and elevation rotations and their driving units, feed positioning systems, protective sentinel systems etc. are inside the shell. They receive commands from a computer located in the observatory building. The transmission of commands and LO signals and reception of IF signals from the antennas is done using one or two optical fibers. Optical modulator and demodulator are used fro conversion from RF to optical and optical to RF respectively. These units are located at the antenna shell and the observatory building.

The distance from the observatory to an antenna may vary between few hundred meters to several km. Having brought the IF signal from the antennas to the observatory building, they are converted to several base-bands followed by sampling and digitizing. Variable instrumental delay is added for compensating the geometric delay to achieve fringe stop condition. Thereafter the signals goes into correlator or pulsar back-end and finally saved in a data storage system. Signals from all antennas are managed in a similar manner.

Assignment Problems-I

1. Astronomical radio signals may be affected by the noise due to atmospheric absorption. Would you prefer to observe a source at 1.4 GHz at the time it rises or when it is at the zenith? Justify your answer.

2. Using a diagram describe the functional blocks of a radiometer.

3. Explain the following terms as applicable to a radiometer receiver: (*i*) Center frequency, (*ii*) Gain, (*iii*) Band-width, (*iv*) Integration time, and (*v*) Receiver temperature.

4. For a radiometer, explain the terms: (*i*) Sensitivity, and (*ii*) Stability.

5. A receiver consists of a LNA followed by a RF amplifier. The LNA gain is 20 dB and its noise figure as 1.1 dB. The RF amplifier has a gain of 60 dB and a noise figure of 3.5 dB. If the efficiency of the transmission line is unity, find the receiver noise temperature given its physical temperature as 300 K.
Hint: You must convert the noise figures in into noise factors and use:

$$T_R = (F-1) \times T_{R(\mathrm{Phy})} \qquad T_R = \frac{1}{\epsilon_T}\left(T_1 + \frac{T_2}{G_1} + \frac{T_3}{G_1\, G_2} + \ldots\right)$$

Assignment Problems-II

6. With the help of a diagram, briefly explain the total power receiver. Why is it called direct radiometer?

7. Using a diagram briefly explain the Dicke receiver.

8. With the help of a diagram, briefly explain the gain modulated Dicke receiver.

9. Draw a block diagram of a null balancing Dicke receiver and explain its Functionality.

10. With the help of a diagram, briefly explain the Graham's receiver.

11. Draw a block diagram of a correlation receiver and explain its functionality.

12. Describe the functionality of an additive interferometer receiver using a block diagram.

Assignment Problems-III

13. Draw the block diagram of a multiplicative interferometer receiver and explain it in details.

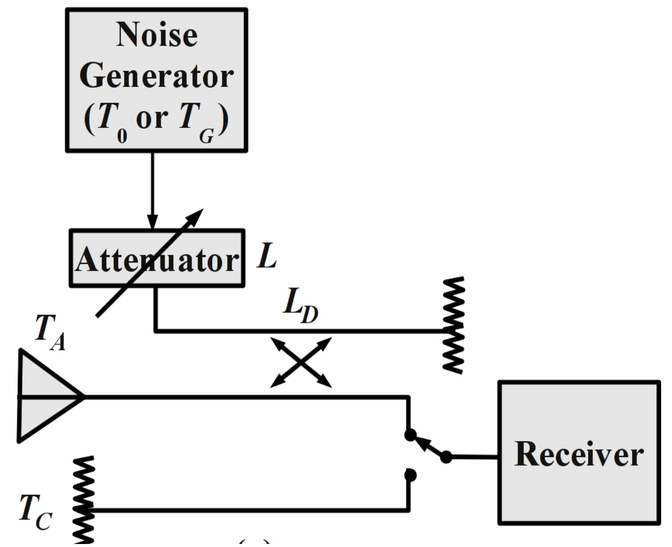
14. Explain the functional blocks of a phase switched receiver using a diagram.

15. An LNA is kept at room temperature (300 K). Its input is connected to a matched load. The output power of the LNA is measured using a RF power meter. When the load is at room temperature, the power meter reading is 1 unit. When it is dipped in liquid nitrogen (77 K), the power meter reads 0.40 unit. Find the receiver temperature. Also calculate the noise figure of the LNA.

Hint:

$$T_R = \frac{P_{\text{off}} T_G - P_{\text{on}} T_0}{P_{\text{on}} - P_{\text{off}}} \qquad T_R = (F-1) \times T_{R(\text{Phy})} \text{ K}$$

16. A noise generator is used in receiver calibration using a directional coupler and an attenuator. It has a fired temperature of 9900 K. If the directional coupler has 20 dB coupling, find the required attenuation to obtain a calibration signal temperature of 10±0.05 K.



Hint:

$$T_{\text{excess}} = \frac{1}{L_D}\left[\frac{T_G}{L} + \left(1 - \frac{1}{L}\right) T_0\right] - \frac{T_0}{L_D} = \frac{T_G - T_0}{L\, L_D}$$

Assignment Problems-IV

17. A point source produces a flux-density of 1000 jansky near the aperture of a single polarized radio telescope antenna. If the effective aperture area is 3 m^2, find the calibration temperature.

Hint:

$$T_{\text{cal}} = \frac{S\,A_e}{2k}$$

18. What is a frequency-switched radiometer and where it is used? Explain Briefly.

19. Using a block diagram explain the operation of a basic pulsar receiver.

20. What is an analog bank spectrometer? Explain with a block diagram.

21. Explain the functionality of a correlation spectrometer using a diagram.

22. What is the difference between in-coherent and coherent de-dispersion in pulsar receives?

23. Draw a generalized block layout of the complete receiving and control chain from observatory building to antennas of a radio array.

THANK YOU