CHAPTER VII

CONCLUSION

In this thesis we have described a 3.5 km interferometer, set up at Ootacamund, for operation at 327 MHz. The Ooty telescope is used as the main element of the interferometer. The electronics for the radio link of the north 13.5 m dish and receiver system for both the north dish and the nearby dish have been completed. Observations of the planet Jupiter, Sgr A and a few Markarian galaxies are described.

In building the interferometer we have taken advantage of the fact that the Ooty telescope has a long north-south aperture, a large collecting area and a $9\frac{1}{2}$ hour tracking capability. This makes it possible to build an interferometer for synthesizing large aperture by putting several small sized antennas at fixed locations around the Ooty telescope. In the first instance, we have constructed a four-element interferometer by putting three parabolic dishes around the Ooty telescope. The interferometer provides a beam of 40" arc in east-west, 3.5' arc in north-south at declination of zero degree and has a confusion limit of about 25 mJy. Though the existing interferometer has a limited (u,v) coverage it has a low confusion limit and therefore it is well suited for detection of weak radio sources. Besides, it is a useful instrument for the study of those sources which have almost line brightness distribution and also radio sources at low declinations.

For generating coherent local oscillator signals at element antennas of the interferometer a phase stable radio link based on the sum frequency principle has been built. The radio link works for a one-way path loss of 75 dB. The IF signals from the remote antennas are brought on microwave links.

The interferometer receivers have system temperatures of about 300°K. The correlation receivers use acoustic surface-wave delay lines which have a bandwidth of 2 MHz. In a single $9\frac{1}{2}$ hour observing period the interferometer, consisting of the Ooty telescope and one 13.5 m dish, for a 30 mJy source provides a signal to noise ratio of 5. The rms phase stability of the entire system is about 12°.

The present on-line computer program for delay control and data acquisition allows observations of only one baseline at a time. Therefore, the observations of Jupiter were made with the interferometer consisting of the Ooty telescope and the 13.5 m north dish. The observations were mainly in the hour-angle range of 00^{h} to $+05^{h}30^{m}$. We have obtained strip brightness distributions along the magnetic equator of Jupiter for the three CML ranges, namely 315° to 75°, 75° to 195° and 195° to 315°, at 327 MHz with a resolution of about 50" arc. Jupiter has been completely resolved along its magnetic equator with two strong peaks those sources which have almost line brightness distribution and also radio sources at low declinations.

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Because of the limited (u,v) coverage the synthesized beam of the Ooty interferometer has large sidelobes, especially at higher declinations. By putting about 20 to 30 small sized antennas as proposed by Swarup and Bagri (1973) it is possible to make an aperture synthesis telescope having a narrow beamwidth of about 15" arc.

However, even to improve the (u,v) coverage of the present interferometer system for a resolution of 40" at least a few more additional antennas are required. In about a year's time it is planned to add two more dishes at intermediate positions to provide sufficient coverage for the low spatial frequency components which will provide a better beam to study the complex structure of radio sources.

APPENDIX 1

GEOMETRICAL RELATIONS FOR A TWO-ANTENNA INTERFEROMETER

In this Appendix we derive the geometrical relations for an interferometer consisting of two antennas P and Q. Let us define the baseline PQ, expressed in RF wavelength λ, as $\overline{B} = \overline{PQ}/\lambda$ and 'h' and 'd' as the hour-angle and declination respectively of the pole of the baseline, X , as shown in Fig.Al.1. Let $S_o(H_o, \delta_o)$ be the direction of the centre of the field of view under observation. Here H_o and δ_o are hour-angle and declination respectively of point S_o on the celestial sphere.
In Fig.Al.1, θ_o is the angle between the source direction $\overline{PS_o}$ and the baseline $\overline{PQ}$ and N is the celestial northpole. Now applying the cosine rule in the spherical triangle NS_oX, we get

$$\cos \theta_o = \sin d \sin \delta_o + \cos d \cos \delta_o \cos(H_o - h) . \qquad (A1.1)$$

The effective baseline for the source at S_o will be $B \sin \theta_o$. We use a Cartesian coordinate system (x,y,z) where z axis is in the direction of the centre of the field of view, S_o, and x and y axes are such that their projections on ground are in the east and north directions respectively. The components of the effective baseline in the x and y directions are the components of the spatial

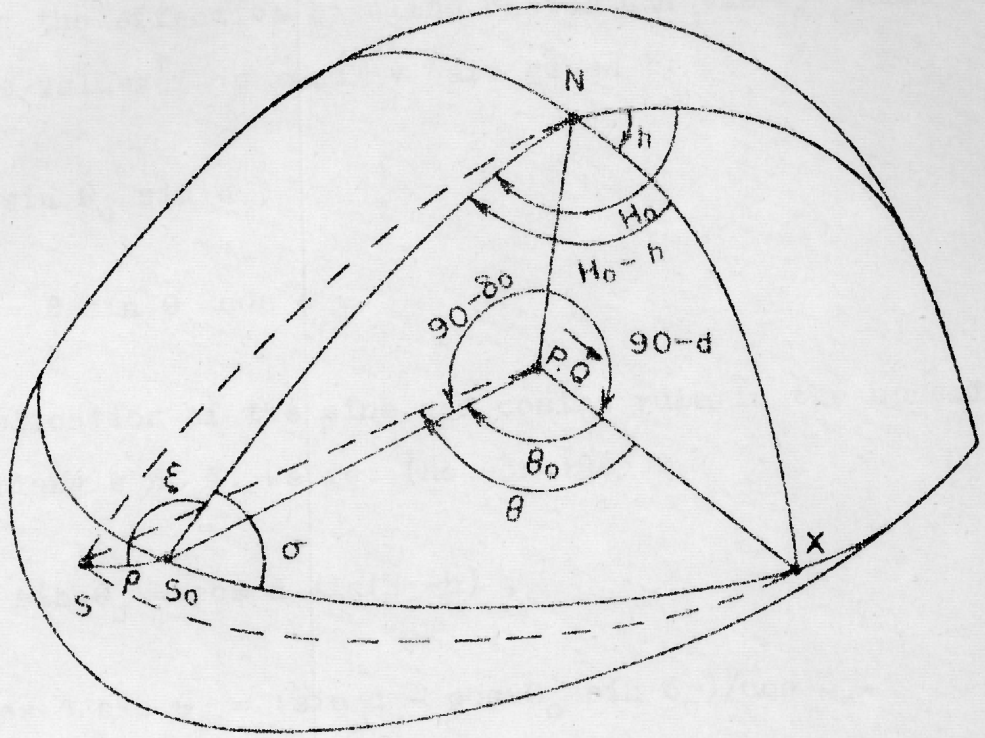

FIG. A1. 1. THE SPHERICAL TRIANGLE SHOWING RELATIONS BETWEEN THE POSITIONS OF THE SOURCE S AND THE BASELINE $\vec{PQ}$ MEETING CELESTIAL SPHERE AT X. N IS THE NORTH POLE. DIRECTION OF THE CENTRE OF THE FIELD OF VIEW UNDER OBSERVATION IS S_0 ($SS_0 = \rho$ AND ANGLE $NS_0S = \xi$)

frequency called u and v respectively. If the angle between the effective baseline vector and the y axis is σ, the values of u and v are given by

$$u = B \sin \theta_o \sin \sigma , \tag{A1.2}$$

$$\text{and } v = B \sin \theta_o \cos \sigma . \tag{A1.3}$$

By application of the sine and cosine rules in the spherical triangle NS_oX, we get (Rowson 1963)

$$\sin \sigma \sin \theta_o = \cos d \sin(H_o - h) , \tag{A1.4}$$

$$\text{and } \cos \sigma \sin \theta_o = (\sin d - \cos \theta_o \sin \delta_o)/\cos \delta_o . \tag{A1.5}$$

Substituting for $\cos \theta_o$ from Eq.(A1.1) in Eq.(A1.5) we get

$$\cos \sigma \sin \theta_o = \sin d \cos \delta_o - \cos d \sin \delta_o \cos(H_o - h). \tag{A1.6}$$

From Fig.A1.1, it is clear that the angle which the baseline $\overline{PQ}$ makes with the axis of rotation of the earth is $(\pi/2-d)$. Therefore, the components of the baseline parallel and perpendicular to the axis of the earth denoted by B_1 and B_2 respectively are given by

$$B_1 = B \sin d , \tag{A1.7}$$

and $B_2 = B \cos d$. (A1.8)

Using Eqs.(A1.4) to (A1.8) in (A1.2) and (A1.3) we get

$$u = B_2 \sin (H_0 - h) \ , \qquad \text{(A1.9)}$$

$$\text{and } v = B_1 \cos \delta_0 - B_2 \sin \delta_0 \cos(H_0 - h) \ . \qquad \text{(A1.10)}$$

For a point source at S_0, a plane wave will arrive at antenna P, by τ_{S_0} seconds later than at antenna Q (see Fig.A1.1) where

$$\tau_{S_0} = (\lambda/c) B \cos \theta_0 \text{ seconds.} \qquad \text{(A1.11)}$$

Now we will consider a point source at $S(H,\delta)$ close to $S_0(H_0,\delta_0)$ such that S is ρ radians away from S_0 and S_0S makes an angle ξ with the direction of the north celestial pole as shown in Fig.A1.1. For a Cartesian coordinate system with direction cosines (ℓ,m,n) with n along z axis PS_0 and ℓ,m along x and y axes with their projections on the ground in the east and north directions respectively, from Fig.A.1, we have (Brouw 1968)

$$\ell = -\sin \rho \sin \xi \ , \qquad \text{(A1.12)}$$

$$\text{and } m = \sin \rho \cos \xi . \qquad \text{(A1.13)}$$

From spherical triangle NSS_o, by applying the sine rule we get

$$\ell = -\sin\rho\sin\xi = \cos\delta\sin(H-H_o)\ , \tag{A1.14}$$

and by applying the cosine rule twice, we get

$$m = \sin\rho\cos\xi = \sin\delta\cos\delta_o - \cos\delta\sin\delta_o\cos(H-H_o) \tag{A1.15}$$

If S is close to S_o such that $(\delta-\delta_o)$ and $(H-H_o)$ are small we can approximate

$$\ell \simeq \cos\delta.\ (H-H_o) = \Delta\alpha'(\text{say})\ , \tag{A1.16}$$

$$m \simeq \sin\delta\cos\delta_o - \cos\delta\sin\delta_o = \sin(\delta-\delta_o) \simeq \delta-\delta_o = \Delta\delta\ . \tag{A1.17}$$

The time delay for signal to arrive at P with respect to Q from S will be

$$\tau_S = \frac{\lambda}{c}\, B\cos\theta\ ,$$

where θ = angle made by the direction of the source S with the baseline $\overline{PQ}$. Thus the difference in delay between signals from S and S_o

$$\tau_S - \tau_{S_o} = \lambda/c\ .\ (B\cos\theta - B\cos\theta_o)\ , \tag{A1.18}$$

From triangle SS_oX,

$$\cos\theta = \cos\theta_o \cos\rho + \sin\theta_o \sin\rho \cos(\sigma+\xi),$$

$$\cos\theta - \cos\theta_o = \cos\theta_o(\cos\rho-1) + \sin\theta_o \sin\rho$$

$$[\cos\sigma \cos\xi - \sin\sigma \sin\xi] \ . \qquad \text{(A1.19)}$$

Using Eqs.(A1.2), (A1.3), (A1.12), (A1.13) and (A1.19) in Eq.(A1.18) we get

$$\tau_S - \tau_{S_o} = \frac{\lambda}{c} \{B[(\cos\rho-1)\cos\theta_o] + u\ell + vm\} \ . \qquad \text{(A1.20)}$$

If ρ is small, $\cos\rho \sim 1$ and we get

$$\tau_S - \tau_{S_o} = \frac{\lambda}{c}[u\ell + vm] \ . \qquad \text{(A1.21)}$$

For a baseline of 4000 λ and for a source 10' away from the centre of the field of view ($\rho \sim 10'$), the maximum phase error due to the assumption $\cos\rho = 1$ is only 4^o. For larger value of ρ this correction may become appreciable. As the Ooty interferometer has about $2^o.5$ field of view in the east-west direction and has a baseline of about 4000 λ, this error may become appreciable for sources away from centre of field of view. With the approximations in (A1.16) and (A1.17) we can rewrite Eq.(A1.21) as

$$\tau_S - \tau_{S_o} = \lambda/c\ [\ u\ \Delta\alpha' + v\ \Delta\delta\ ]\ . \qquad \text{(A1.22)}$$

APPENDIX 2

BASELINE PARAMETERS OF A TWO-ANTENNA INTERFEROMETER

In this Appendix we derive expressions for the baseline parameters B_1, B_2 and h in terms of the measured coordinates of the antennas. Here B_1 and B_2 are components of the baseline $\bar{B}$ parallel and perpendicular to the axis of rotation of the earth respectively and expressed in units of RF wavelength and h is hour-angle of the pole of the baseline $\bar{B}$. The vector $\bar{B}$ is formed by the line joining the two antennas $P(x_1, y_1, p_1)$ and $Q(x_2, y_2, p_2)$. Cartesian coordinate system (x,y,p) with axes along east, north and zenith respectively is used in the survey of the location of the antennas.

Let us define N as the north celestial pole, z as the zenith direction, X as the direction of the pole of the baseline $\overline{PQ}$ and φ as the geographic latitude of antenna P , as shown in Fig. A2.1. We first find values of the azimuth angle A and zenith angle z of the baseline $\overline{PQ}$. These are given by

$$\tan A = (x_2-x_1)/(y_2-y_1) , \tag{A2.1}$$

and
$$\tan z = [(x_2-x_1)^2 + (y_2-y_1)^2]^{\frac{1}{2}}/p_2-p_1 . \tag{A2.2}$$

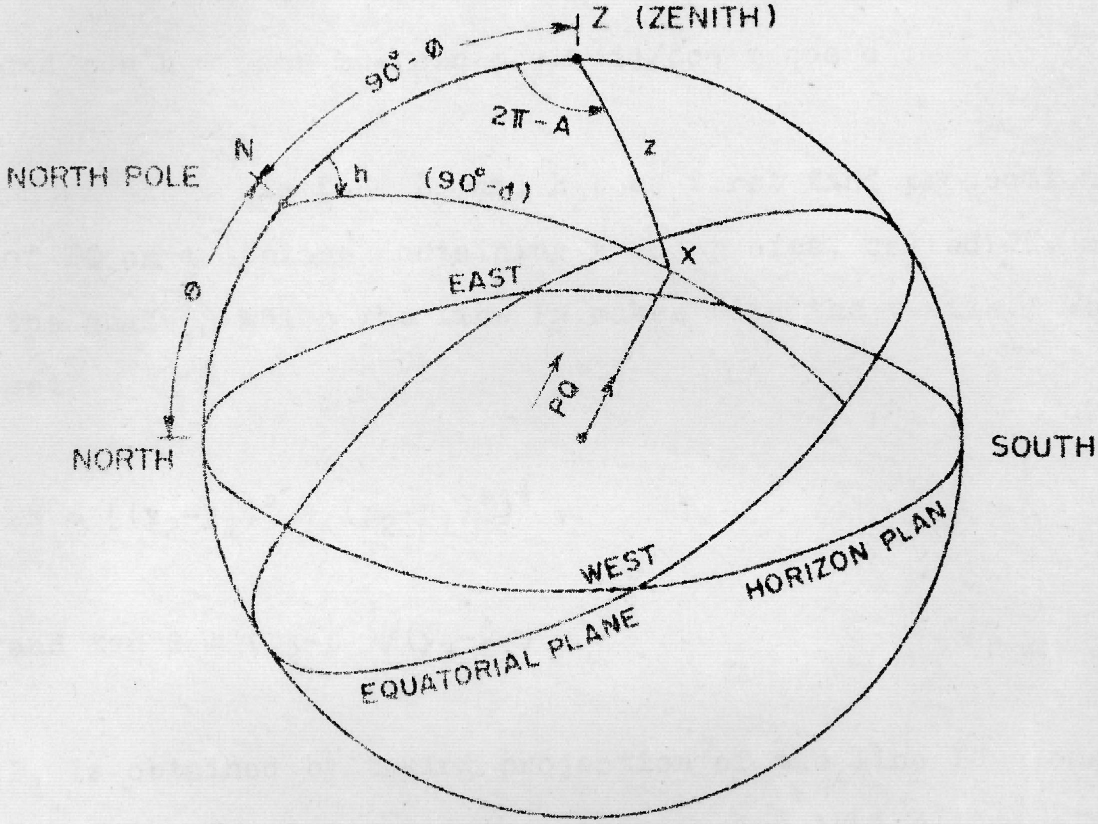

FIG. A2·1 SPHERICAL TRIANGLE SHOWING RELATION BETWEEN EQUATORIAL COORDINATES AND AZIMUTH-ZENITH ANGLES

From the spherical triangle NZX, the declination d and hour-angle h of the pole X are given by

$$\sin d = \sin \varphi \cos z + \cos \varphi \sin z \cos A , \tag{A2.3}$$

$$\text{and } \cos h = (\cos z - \sin \varphi \sin d)/\cos \varphi \cos d . \tag{A2.4}$$

To find B_1 and B_2, we first find projection of PQ on the plane containing y and p axes, called PN, and the angle β which the line PN makes with the y-axis. We get

$$PN = [(y_2-y_1)^2 + (p_2-p_1)^2]^{\frac{1}{2}} , \tag{A2.5}$$

$$\text{and } \tan \beta = (p_2-p_1)/(y_2-y_1) . \tag{A2.6}$$

B_1 is obtained by taking projection of the line PN along the axis of rotation of the earth, which is inclined at an angle φ to the y-axis. Thus, we get

$$B_1 = \lambda^{-1}[(PN \cos(\beta-\varphi)] . \tag{A2.7}$$

B_2 is obtained by taking the vector sum of the component of PN perpendicular to the axis of rotation of the earth and the x-component of the line $\overline{PQ}$. Thus, we get

$$B_2 = \lambda^{-1}. [(PN)^2 \sin^2 (\beta-\varphi) + (x_2-x_1)^2]^{\frac{1}{2}} . \tag{A2.8}$$

APPENDIX 3

ALLOWABLE ONE-WAY PATH LOSS FOR A CLOSED LOOP SYSTEM

In this Appendix we briefly review the limitations on the allowable path attenuation for a closed loop system in which the reference or correction signal at any point along the path is formed by combining the two signals travelling in opposite directions. The permissible attenuation depends upon the allowable phase error of the resultant reference or correction signal. Here we calculate the phase errors that arise due to the receiver and oscillator noise. The phase errors due to the presence of reflections on the transmission path have been discussed by Swarup and Yang(1961) and Jones (1972). The contribution by oscillator noise is particularly influenced by the effect of the reflection coefficient ρ and the use of a multiplier which has a finite value of the proportionality constant ε . ε is defined as the ratio of the amplitudes of second harmonics, due to signal at any port, to the product of the two input signals.

Let S_1 at frequency $(\omega_0/2 + \omega_a)$ and S_2 at frequency $(\omega_0/2)$ be the power outputs of the two oscillators at home and remote stations respectively; N_1 and N_2 be their noise power density per Hz, kT per Hz be the receiver noise, and β_1 & β_3 be the fractions of power coupled to the multiplier from the oscillator at home station and the signal αS_2 coming from remote station respectively. The signals at the two inputs of the multiplier are X and Y. These are shown in Fig.A3.1. Also, let B_1, B_2 and B_3 be the equivalent bandwidths in Hz of the receiver IF, output filter and the oscillator noise respectively, as shown in Fig.A3.2. The output of the multiplier is given by

$$V_o = XY + \varepsilon\,(X^2+Y^2). \tag{A3.1}$$

The desired reference signal at frequency $(\omega_o+\omega_a)$ in case of the sum frequency system, and at ω_a in case of the difference frequency system, is formed by multiplying signals $\beta_1 S_1$ from home station and $\alpha\beta_3 S_2$ from remote station. This can be written as

$$S = (\alpha\beta_3\beta_1 S_1 S_2)^{\frac{1}{2}}. \tag{A3.2}$$

The phase errors are caused by the noise components present in the output of the multiplier over the bandwidth B_2. The noise terms are produced by the multiplication of the individual terms of the inputs X and Y (excluding signal S defined above), and ε times the second harmonics of the two inputs. We assume that the signal $\beta_1 S_1$ from home station is strong such that

(i) the product due to $\beta_1 N_1$ and $\rho^2\beta_3 N_1$ over B_2 is larger than the product due to $\beta_1 N_1$ and $\rho^2\beta_3 N_1$ over B_3.i.e. $(2N_1 B_2 S_1)^{\frac{1}{2}} > N\,B_3$ and

(ii) the product due to $\beta_1 S_1$ and kT/Hz over B_2 is larger

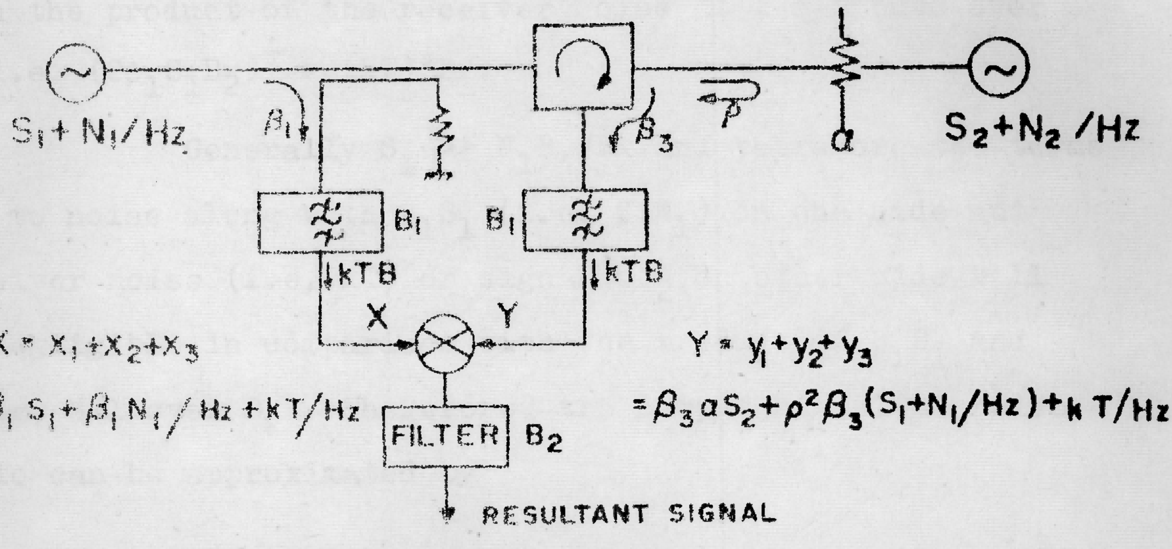

FIG. A3.1 BLOCK DIAGRAM SHOWING SIGNAL AND NOISE COMPONENTS AT VARIOUS STAGES FOR CALCULATING ALLOWABLE ONE WAY PATH LOSS FOR A CLOSED LOOP SYSTEM

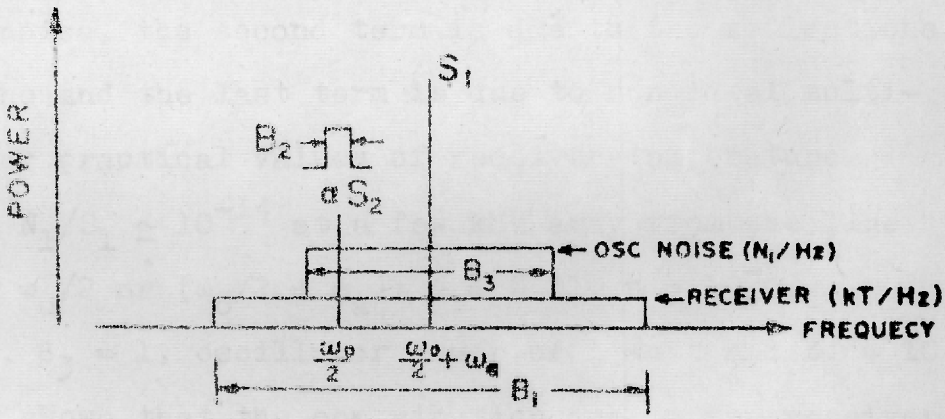

FIG. A3.2 SPECTRAL DISTRIBUTION OF VARIOUS SIGNAL AND NOISE COMPONENTS than the product of the receiver noise at the inputs over B_1 i.e. $(2\beta_1 S_1 B_2)^{\frac{1}{2}} > (kT)^{\frac{1}{2}} B_1$.

Generally $S_1 >> N_1 B_3/B_2$ and therefore the terms due to noise along with $\beta_1 S_1$ (i.e. $\beta_1 N_1$) on one side and receiver noise (i.e. kT) or signal αS_2 on other side will be negligible in comparison with the product of $\beta_1 S_1$ and kT/Hz, all over B_2. Therefore, the output signal to noise ratio can be approximated by

$$\frac{S}{N} \simeq \frac{(\alpha S_2 \beta_3)^{\frac{1}{2}}}{(2kTB_2)^{\frac{1}{2}} + (4\beta_3 \rho^2 N_1 B_2)^{\frac{1}{2}} + (4\varepsilon\beta_1 N_1 B_2)^{\frac{1}{2}}} . \qquad (A3.3)$$

Here in the denominator the first term is due to the receiver noise, the second term is due to the reflections on the line and the last term is due to non ideal multiplier. For practical values of receiver temperature $\lesssim 10^3$ $^{\circ}$K, $N_1/S_1 \simeq 10^{-14}$ at a few kHz away from the line frequency $\omega_o/2$ or $(\omega_o/2 + \omega_a)$; $\rho = 0.01$, $\varepsilon = 10^{-3}$, $\beta_1 = 10^{-3}$, $\beta_3 = 1$, oscillator power of 1 Watt and $B_2 = 10$ Hz, it can be shown that the contribution due to the receiver noise is much smaller than that due to the reflections and the non ideal behaviour of the multiplier. For a rms phase error of 6° the output signal to noise ratio should be greater than 20 dB, which implies that

$$\frac{S}{N} \simeq \frac{(\alpha S_2\ \beta_3)^{\frac{1}{2}}}{(4\beta_3 \rho^2 N_1 B_2)^{\frac{1}{2}} + (4\varepsilon\beta_1 N_1 B_2)^{\frac{1}{2}}} > 10 \quad . \qquad (A3.4)$$

When two oscillators are used at the two ends of the transmission path the maximum loss for the above assumed values works out to be about 140 dB. However, when the signal from the remote end is formed by reflection of the incident signal (reflected modulation technique) the maximum allowable one-way path loss has to be limited to about 70 dB. These limits are mainly due to reflections on the transmission path, the non ideal behaviour of the multiplier and presence of noise in the oscillator output.

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