Phase Transfer via AUSSAT

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The possibility of transferring phase between telescopes via AUSSAT has been examined in AT/10.3/001 with the conclusion that a viable system can be implemented. Some aspects which were not considered in any detail in the earlier report are examined here and a practical implementation is suggested. A system linking 6 telescopes could be run continuously at an annual rental of about $20 K/yr for the use of AUSSAT plus the Federal licence fee for each ground terminal.

Phase transmission by the two-way, tone-difference method is assumed. The tones of a pair are separated by $\geq$192 MHz and the tone pairs are displaced at regular intervals of 4 KHz with respect to each other. Figure 1 is a sketch showing the transmitted spectra at each ground station.

The main analyses are found in Appendices 1 to 11. Substantial portions of these are derived from van Ardenne et al. (1983 and 1981), van Ardenne (1983), as well as AT/10.3/001. Other useful information was found in Feher (1983), AT/10.3/004 and the AUSSAT Network Designers Guide.

The main conclusions are as follows:

1) A phase transfer uncertainty equivalent to $\Delta\tau \leq 2.3$ picosecond rms can be achieved in a meaurement time of 1 second with an overall signal-to-noise ratio of a tone of 48.1 dB Hz and a tone separation $\geq$192 MHz. This uncertainty would reduce fringe visibility by 5% at an observing frequency of 22 GHz.

2) Astronomical data needs to be phase corrected every second since independent rubidium standards only achieve a relative stability as good as this on time scales $\leq$ 1 second.

3) Careful design and environmental control can keep $\Delta\dot{\tau} \leq 10^{-14}$ so that data can be integrated coherently for up to a few hundred seconds within whatever limitations are imposed on the astronomical signals by the ionosphere and the troposphere.

4) Channels supporting digital communications at a few Kbs^{-1} can be implemented using similar "carrier" powers to those required by the tones.

5) Up to 6 telescopes can be supported within the power and bandwidth restrictions of single-channel-per-carrier useage of a transponder. Two of these channels are required separated by $\geq$192 MHz. More telescopes could be added as necessary.

References

van Ardenne, O'Sullivan, and Dianous, IEEE TRANS. IM-32, pp 370-376, 1983.

van Ardenne, O'Sullivan, and Buiter, NFRA Note 341, 1981.

van Ardenne, NRFA Note 395, 1983.

Feher, "Digital Communications: Satellite/Earth Station Engineering", Prentice Hall 1983.

Figures

1. The transmitted spectra at each ground station.

2. The phase transfer uncertainty of an AUSSAT link (expressed in picoseconds) as a function of averaging time compared with the uncertainty between two independent rubidium standards.

3. Schematic diagram of a processing system for the tones.

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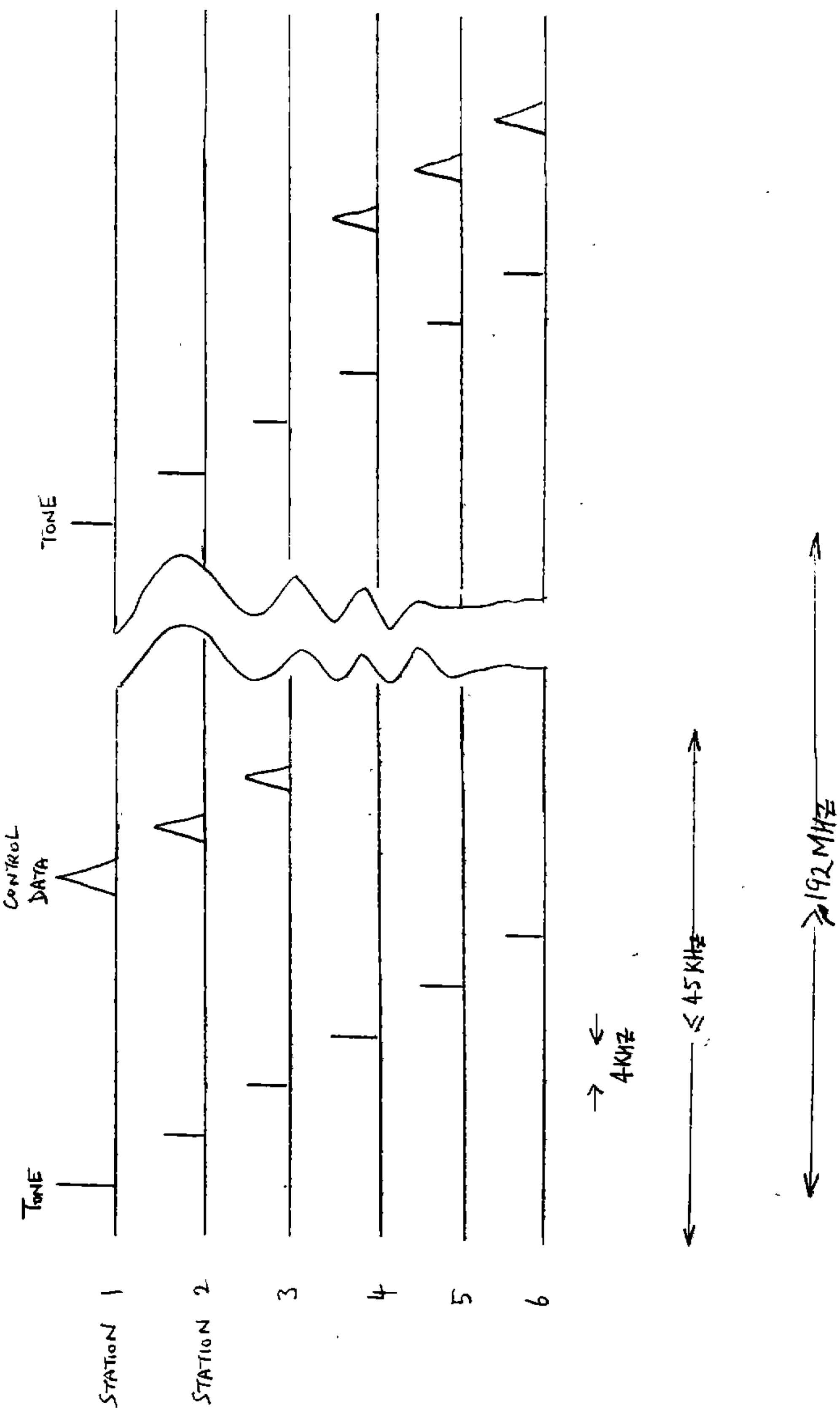

TONE
CONTROL DATA
TONE
STATION 1
STATION 2
3
4
5
6
4KHz
≤ 45 KHz
≥ 192 MHz

Figure 1

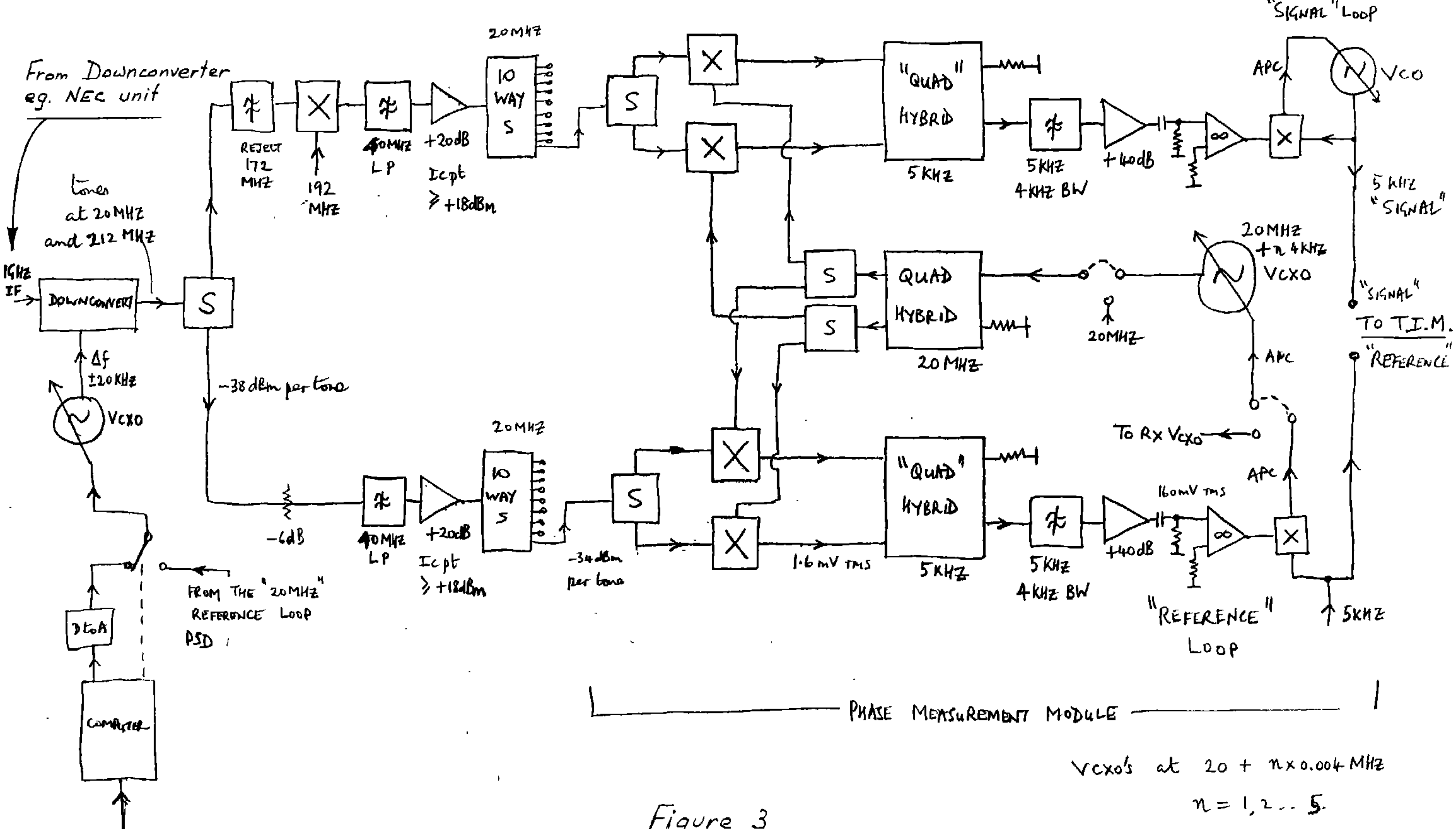
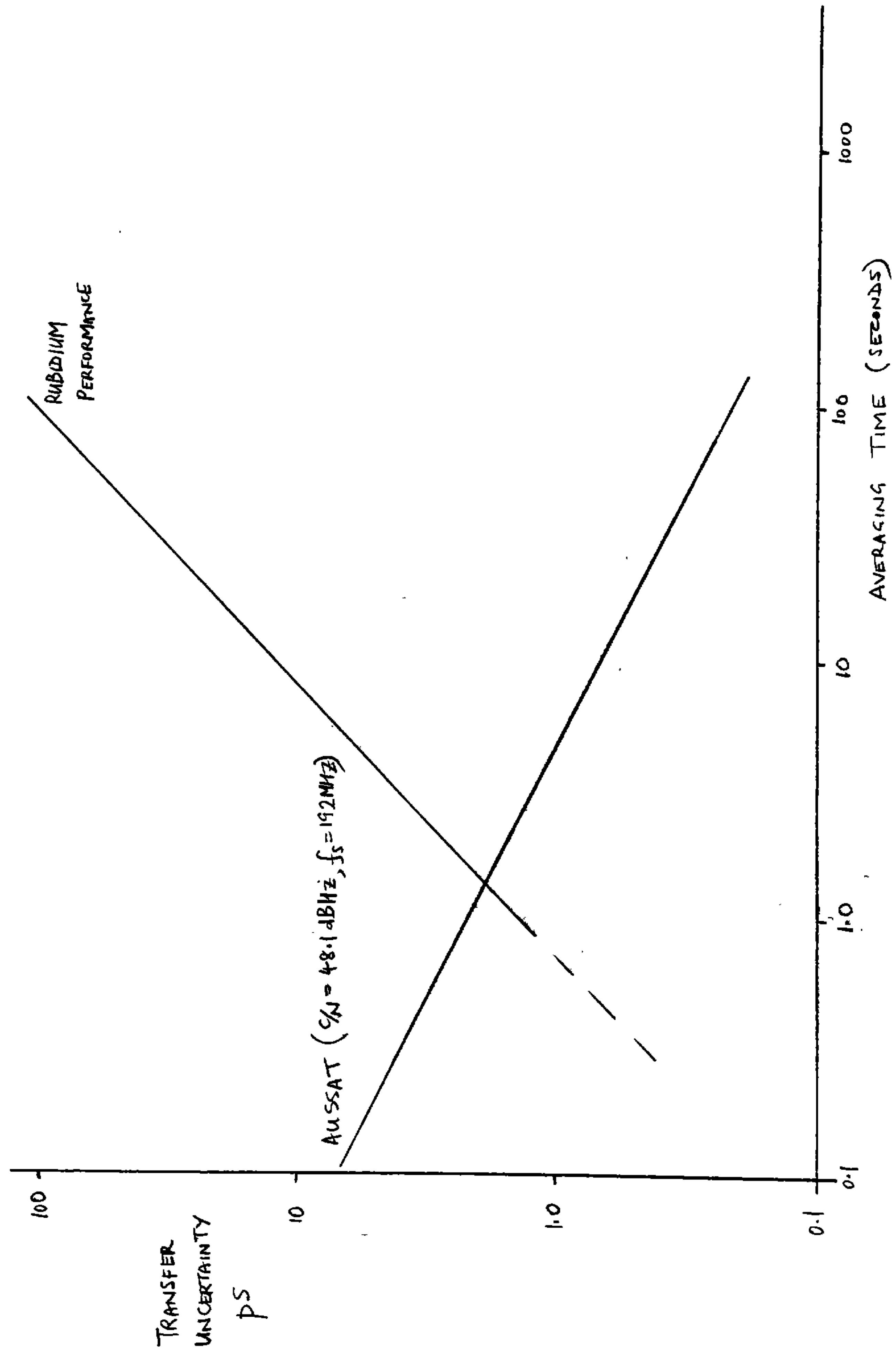

From Downconverter eg. NEC unit
1GHz IF
DOWNCONVERT
tones at 20MHz and 212 MHz
Δf ±20KHz
VCXO
D to A
COMPUTER
FROM THE "20MHz" REFERENCE LOOP PSD
S
REJECT 172 MHz
192 MHz
40MHz LP
+20dB
Icpt ≥ +18dBm
20MHz
10 WAY S
-38 dBm per tone
-6dB
40MHz LP
+20dB
Icpt ≥ +18dBm
20MHz
10 WAY S
-34 dBm per tone
"QUAD" HYBRID
5KHz
QUAD HYBRID
20 MHz
"QUAD" HYBRID
5KHz
1.6mV rms
5KHz 4KHz BW
+40dB
5KHz 4KHz BW
+40dB
160mV rms
APC
"SIGNAL" LOOP
VCO
5 kHz "SIGNAL"
20MHz +n 4KHz VCXO
20MHz
APC
To Rx VCXO
"SIGNAL"
TO T.I.M.
"REFERENCE"
"REFERENCE" LOOP
5KHz
PHASE MEASUREMENT MODULE
Vcxo's at 20 + n x 0.004 MHz
n = 1, 2 .. 5.

Figure 3



RUBIDIUM PERFORMANCE
AUSSAT (C/N = 48.1 dBHz, fs = 192 MHz)
TRANSFER UNCERTAINTY ps
100
10
1.0
0.1
0.1
1.0
10
100
1000
AVERAGING TIME (SECONDS)

Figure 2

Appendices

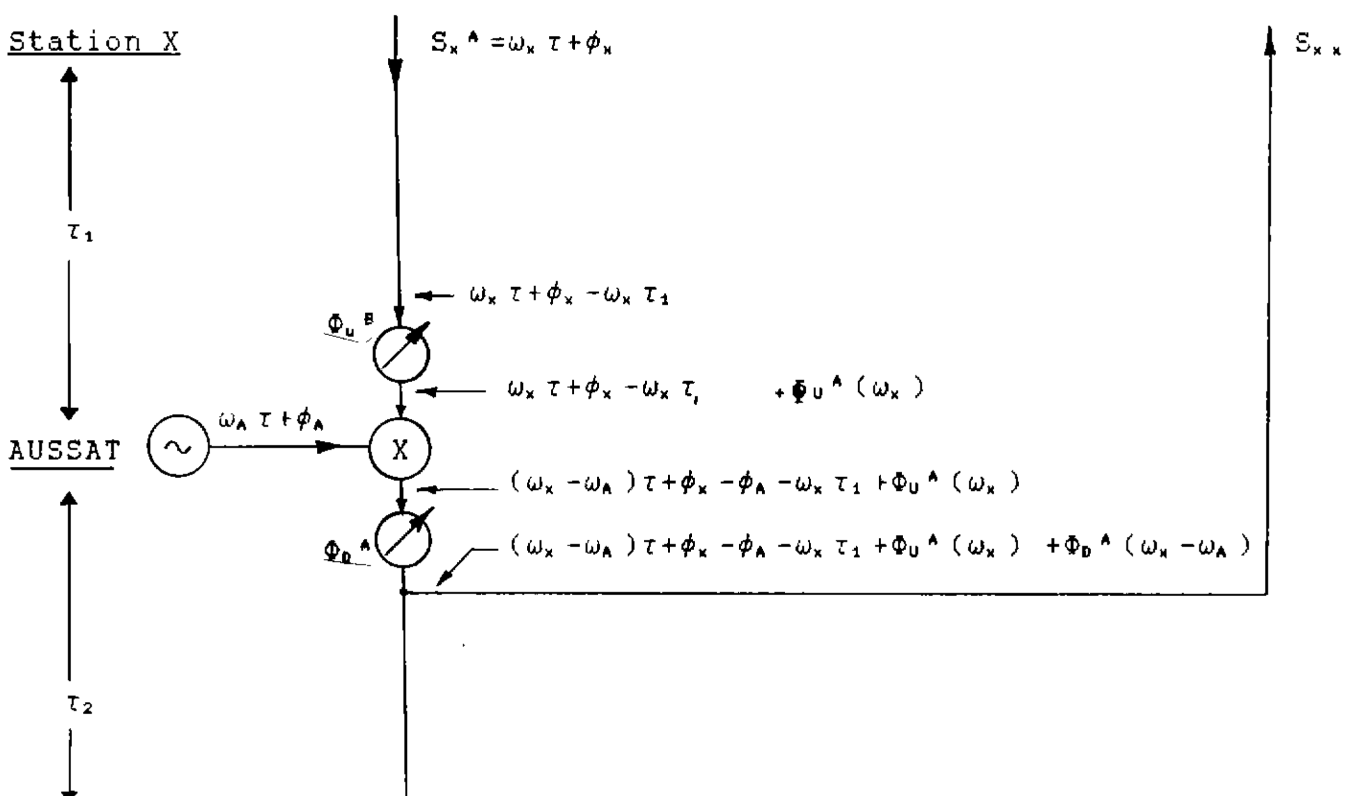
A1. Phase transfer between station X and station Y via AUSSAT

The following notation will be used.

$S_{xy}^{A} = \omega\tau + \phi$ — where S represents the phase of a signal transmitted from X to Y

through transponder A. The angular frequency of the signal is ω radians per second, and the phase angle is ϕ radians.

$\Phi_U^{A}(\omega)$, $\Phi_D^{A}(\omega)$ — represent phase shifts in transponder A in U the uplink and D the downlink at frequency ω



$$S_{xy}^{A} = (\omega_x - \omega_A)\tau + \phi_x - \phi_A - \omega_x(\tau_1 + \tau_2) + \omega_A\tau_2 + \Phi_U^{A}(\omega_x) + \Phi_D^{A}(\omega_x - \omega_A)$$

and

$$S_{xx}^{A} = (\omega_x - \omega_A)\tau + \phi_x - \phi_A - 2\omega_x\tau_1 + \omega_A\tau_1 + \Phi_U^{A}(\omega_x) + \Phi_d^{A}(\omega_x - \omega_A)$$

Similarly for $S_x^B = (\omega_x + \omega_s)\tau + \phi_x + \phi_s^x$ where S represents values for the standard frequencies to be transferred.

Thus $S_x^B \rightarrow S_{xy}^B = (\omega_x + \omega_s - \omega_B)\tau + \phi_x + \phi_s^x - \phi B$

$$- (\omega_x + \omega_s)(\tau_1 + \tau_2) + \omega_B \tau_2 + \Phi_U^B(\omega_x + \omega_s)$$

$$+ \Phi_D^B(\omega_x + \omega_s - \omega_B)$$

and $S_{xx}^B = (\omega_x + \omega_s - \omega_B)\tau + \phi_x + \phi_s^x - \phi_B$

$$- 2(\omega_x + \omega_s)\tau_1 + \omega_B \tau_1 + \Phi_u^B(\omega_x + \omega_s)$$

$$+ \Phi_D^B(\omega_x + \omega_s - \omega_B)$$

Similar relations exist for the transmissions from Y to X, viz.

$S_Y^A = \omega_y \tau + \phi_y \rightarrow S_{yx}^A$ and S_{yy}^A

and $S_y^B = (\omega_y + \omega_s)\tau + \phi_y + \phi_s^y \rightarrow S_{yx}^B$ and S_{yy}^B

At Y, measure $R_{xy} = S_{xy}^B - S_{xy}^A - (S_y^B - S_y^A) = (\omega_A - \omega_B)\tau + \phi_s^x - \phi_s^y$

$$+ (\phi_A - \phi_B) - \omega_s(\tau_1 + \tau_2)$$

$$- (\omega_A - \omega_B)\tau_2 + \Phi_u^B(\omega_x + \omega_s) + \Phi_D^B(\omega_x + \omega_s - \omega_B)$$

$$- \Phi_u^A(\omega_x) - \Phi_D^A(\omega_x - \omega_A)$$

and $R_{yy} = S_{yy}^B - S_{yy}^A - (S_y^B - S_y^A) = (\omega_A - \omega_B)\tau + \phi_A - \phi_B$

$$- 2\omega_s \tau_2 - (\omega_A - \omega_B)\tau_2$$

$$+ \Phi_u^B(\omega_y + \omega_s) + \Phi_d^B(\omega_y + \omega_s - \omega_B) -$$

$$- \Phi_u^A(\omega_y) - \Phi_D^A(\omega_y - \omega_A)$$

At X, measure $R_{yx} = S_{yx}^B - S_{yx}^A - (S_x^B - S_x^A) = (\omega_A - \omega_B)\tau + \phi_s^y - \phi_s^x$

$$+ (\phi_A - \phi_B) - \omega_s(\tau_1 + \tau_2)$$

$$- (\omega_A - \omega_B)\tau_1 + \Phi_u^B(\omega_y + \omega_s) + \Phi_D^B(\omega_y + \omega_s - \omega_B)$$

$$- \Phi_u^A(\omega_y) - \Phi_D^A(\omega_y - \omega_A)$$

and $R_{xx} = S_{xx}^B - S_{xx}^A - (S_x^B - S_x^A) = (\omega_A - \omega_B)\tau + \phi_A - \phi_B$

$$- 2\omega_s \tau_1 - (\omega_A - \omega_B)\tau_1$$

$$+ \Phi_u^B(\omega_x + \omega_s) + \Phi_D^B(\omega_x + \omega_s - \omega_B) - \Phi_u^A(\omega_x)$$

$$- \Phi_D^A(\omega_x - \omega_A)$$

$$R_{xy} - R_{yx} = 2(\phi_s^x - \phi_s^y) - (\omega_A - \omega_B)(\tau_2 - \tau_1) + \Phi_u^B(\omega_x + \omega_s)$$

$$- \Phi_u^B(\omega_y + \omega_s)$$

$$+ \Phi_D^B(\omega_x + \omega_s - \omega_B) - \Phi_D^B(\omega_y + \omega_s - \omega_B) - \Phi_u^A(\omega_x)$$

$$+ \Phi_u^A(\omega_y) - \Phi_D^A(\omega_x - \omega_A) + \Phi_D^A(\omega_y - \omega_a)$$

$$= 2(\phi_s^x - \phi_s^y) - (\omega_A - \omega_B)(\tau_2 - \tau_1) + \Delta\Phi_{AUBBAT} \quad .....(1)$$

where $\Delta\Phi_{AUSSAT} \simeq (\omega_x - \omega_y)\{\frac{\delta\Phi_u}{\delta\omega}^B(\omega_y + \omega_s) + \frac{\delta\Phi_D}{\delta\omega}^B(\omega_y + \omega_s - \omega_B)$

$$-\frac{\delta\Phi_u}{\delta\omega}^A(\omega_y) \quad -\frac{\delta\Phi_D}{\delta\omega}^A(\omega_y - \omega_A)\}$$

Also $R_{xx} - R_{yy} = 2\omega_s(\tau_2 - \tau_1) + (\omega_A - \omega_B)(\tau_2 - \tau_1) + \Delta\Phi_{AUSSAT}$ .....(2)

Eliminating $(\tau_2 - \tau_1)$ from (1) and (2) gives

$$\phi_s^x - \phi_s^y = \tfrac{1}{2}[R_{xy} - R_{yx} + (R_{xx} - R_{yy})\left(\frac{\omega_A - \omega_B}{(2\omega_s + \omega_A - \omega_B)}\right) - \Delta\Phi_{AUSSAT}\{1 + \frac{(\omega_A - \omega_B)}{(2\omega_s + \omega_A - \omega_B)}\}]$$

$_B)$ $\{2\omega_s + \omega_A - \omega_B\}$

...... (3)

An examination of this expression shows that there is no need to measure R_{xx} and R_{yy} if $\omega_A = \omega_B$, i.e. if the transponders are coherent. It can also be seen that $\Delta\Phi_{AUSSAT}$ can be minimised by making $(\omega_x - \omega_y)$ as small as possible.

<u>Noise</u>

If $(\omega_A - \omega_B) << 2\omega_s$ then

$$\phi_s^x - \phi_s^y \simeq \tfrac{1}{2}[R_{xy} - R_{yx}]$$

Therefore, the uncertainties in the determinations of the relative phase and in the measurements are related by

$$\Delta\phi_s = \Delta(\phi_s^x - \phi_s^y)$$

$$= \tfrac{1}{2}[\Delta R_{xy} - \Delta R_{yx}]$$

$$\langle\Delta\phi_s^2\rangle = \tfrac{1}{4}\langle[\Delta R_{xy} - \Delta R_{yx}]^2\rangle$$

$$= \tfrac{1}{4}[\langle\Delta R_{xy}^2\rangle + \langle\Delta R_{yx}^2\rangle]$$

if the uncertainties in R_{xy} and R_{yx} are uncorrelated with one another.

Contributions to ΔR_{xy} come from noise added to S_{xy}^B and S_{xy}^A, and to ΔR_{yx} from S_{yx}^B and S_{yx}^A. Therefore

$$\langle\Delta\phi_s^2\rangle = \tfrac{1}{4}\langle(\Delta S_{xy}^B)^2\rangle + \langle(\Delta S_{xy}^A)^2\rangle + \langle(\Delta S_{yx}^B)^2\rangle + \langle(\Delta S_{yx}^A)^2\rangle]$$

if these are independent noise contributions and

$$\langle \Delta\phi_s^2 \rangle = \langle (\Delta S)^2 \rangle$$

if all the signals have the same ratio of signal to noise and, therefore the same mean square phase uncertainty $\langle (\Delta S)^2 \rangle$

While noise is added to the signals in receptions on AUSAT and on the ground. The relation between phase uncertainty and signal-to-noise ratio is

$$\langle (\Delta S)^2 \rangle = \tfrac{1}{2}\left[\frac{S}{N}\right]^{-1} (\text{radians})^2 \text{ provided that } \frac{[S]}{[N]} \gg 1.$$

The ratio is of signal power to noise power for a tone received at a ground station. The equivalent rms uncertainty in time transfer between the standards is

$$\sigma_\tau = \frac{1}{\omega_s}\left[\langle \Delta\phi_s^2 \rangle\right]^{\frac{1}{2}}$$

$$= \frac{1}{2\sqrt{2}\pi f_s}\left[\frac{S}{N}\right]^{-\frac{1}{2}} \text{ seconds} \qquad \ldots\ldots(4)$$

The signal-to-noise ratio depends on the equivalent noise bandwidth of the measuring system.

$$\frac{[S]}{[N]} = \frac{C}{Nob}$$

where C is the average carrier (signal) power, No is the power spectral density of the added noise i.e. the average power in 1 Hz bandwidth, and b is the noise bandwidth of the measuring system. The simplest measuring system just integrates the data for τ_m seconds. The two-sided noise bandwidth of this process is just $b = 1/\tau_m$

$$\frac{[S]}{[N]} = \frac{C\tau_m}{No}$$

and $$\sigma_\tau = \frac{1}{2\sqrt{2}\pi f_s}\left[\frac{C\tau_m}{No}\right]^{\frac{1}{2}}$$

A2. The effect of the ionosphere on phase transfer by AUSSAT

The effect of the free electrons in the ionosphere is to change the electrical length of a path through the ionosphere by

$$l_i \simeq -\frac{40.5}{f^2} \int_0^\infty N_E \, dl$$

where l_i is in metres, f in Hz, and N_E the free electron density in m^{-3}

The phase change due to the ionosphere is

$$\phi_i = \frac{-2\pi l_i}{\lambda}$$

$$= +\frac{2\pi}{C} \times \frac{40.5}{f} \int_0^\infty N_E \, dl$$

$$\simeq +\frac{8.5\times10^{-7}}{f} \text{ TEC sec (Z)}$$

where TEC is the total electron content $\int_0^\infty N_E \, dl$ along a path in the zenith and Z is the zenith angle of the path with respect to the ionosphere.

The relative phase shift of the 2 tones of a transmission is what matters in phase transfer via AUSSAT. This is best expressed as

$\Delta\phi = \omega_s \tau_{Ni}$ where ω_s is the separation of the tones and τ_{Ni} is the group delay of the Nth path through the ionosphere. The group delay contribution due to the ionosphere is

$$\tau_i = -\frac{\delta\phi_i}{\delta\omega} = \frac{1.35\times10^{-7}}{f^2} \text{ TEC sec (Z)}$$

Since it is desirable to separate the pairs of tones by $(\omega_x - \omega_y) < 0.5$ MHz, the frequencies to be substituted in the expression for τ_i can be a generalised uplink frequency $f_u \simeq 14.185$ GHz and a generalised downlink frequency $f_D \simeq 12.437$ GHz.

The effect of the ionosphere on $(R_{xy} - R_{yx})$ is

$$\Delta(R_{xy} - R_{yx}) \simeq -\omega_s [\tau_{1i}(f_u) + \tau_{2i}(f_D) - \tau_{2i}(f_u) - \tau_{1i}(f_D)]$$

$$= +\omega_s [\{\tau_{1i}(f_D) - \tau_{1i}(f_u)\} + \{\tau_{2i}(f_D) - \tau_{2i}(f_u)\}]$$

$$= \omega_s \; 1.35\text{x}10^{-7} [\text{TEC}_1 \, \text{Sec}(Z_1) + \text{TEC}_2 \, \text{Sec}(Z_2)] \left[\frac{1}{f_D^2} - \frac{1}{f_U^2}\right]$$

$$= \omega_s \; 2\text{x}10^{-28} [\text{TEC}_1 \, \text{Sec}(Z_1) + \text{TEC}_2 \, \text{Sec}(Z_2)]$$

$$\simeq \omega_s \; 4\text{x}10^{-28} \; \text{TEC Sec}(Z)$$ since the TEC's and Z's

for the AT are likely to be similar.

The effect on time transfer between the standards is

$$\Delta(\tau_x - \tau_y) = \frac{\Delta R_{xy} - \Delta R_{yx}}{2 \; \omega_s}$$

$$= 2\text{x}10^{-28} \; \text{TEC Sec}(Z_1)$$

$$= 2.7\text{x}10^{-28} \; \text{TEC for } Z \simeq 40^\circ$$

Note that $\Delta(\tau_x - \tau_y)$ is independent of the frequency separation, ω_s, of the tones. The worst case TEC for solar maximum is ~ 10^{18} m^{-3},

$$\text{diurnally} \quad \left[\frac{\delta \text{TEC}}{\delta \tau}\right] \leqslant \frac{10^{18}}{4\text{x}3600} \simeq 7\text{x}10^{13} \; m^{-3} s^{-1}$$

$$\text{and for TID's} \quad \left[\frac{\delta \text{TEC}}{\delta \tau}\right] \simeq \frac{10^{18}}{50} \text{ x } \frac{2\pi}{600} = 2\text{x}10^{14} \; m^{-3} s^{-1}$$ (2% in 10 minutes)

Then $\quad \Delta(\tau_x - \tau_y) \simeq 270$ pS $= \Delta\tau_I$

diurnally $\Delta(\tau_x - \tau_y)/\Delta\tau \simeq 1.9\text{x}10^{-14} = \Delta\dot{\tau}_{ID}$

and for TID's $\Delta(\tau_x - \tau_y)/\Delta\tau \simeq 5.4\text{x}10^{-14} = \Delta\dot{\tau}_{IT}$

These values can be compared with the ionospheric effect on a wavefront from an astronomical object to a telescope, again at a zenith angle of 40°.

Observing Freq. (GHz)	0.4	5	10	22.3	85
$\Delta\tau_I$	1.1 μs	7.05 nS	1762 pS	354 pS	24.4 pS
diurnal $\Delta\dot{\tau}_{ID}$	7.7×10^{-11}	4.9×10^{-13}	1.2×10^{-13}	2.5×10^{-14}	1.7×10^{-15}
TID $\Delta\dot{\tau}_{IT}$	2.2×10^{-10}	1.4×10^{-12}	3.5×10^{-13}	7.1×10^{-14}	4.9×10^{-15}

Of course, some degree of cancellation of the longer term variations can be expected for telescopes only a few hundred kilometres apart, but the effect of TID's will be cumulative. These figures show that phase transfer via AUSSAT will not limit astronomical observations at or below frequencies of 22 GHz provided that the electronics do not contribute larger drift rates than the ionosphere.

A3. The neutral atmosphere

A rough estimate of the magnitudes of the phase errors observed at the VLA for baselines out to 30 km is given in AT/10.3/004, viz.

$$P \simeq f\,(\mathrm{GHz})\, D_{(km)}{}^{0.7} \text{degrees, for time scales}$$

between a few minutes and a few hours. If it is assumed that these phase fluctuations are caused by atmospheric disturbances drifting over the array at the local wind speed, υ, then D can be replaced by $\upsilon\tau$, and the phase converted into an equivalent time delay

$$\Delta\tau_A \simeq 0.11\ \tau^{0.7}\ \mathrm{pS}$$

where a wind velocity of 10 ms^{-1} has been used and τ is the time scale in seconds over which $\Delta\tau_A$ occurs.

Differentiating gives $\Delta\dot{\tau}_A = 0.074\ \tau^{-0.3}$ as a rough estimate of the delay rate. $\Delta\dot{\tau}_A$ is the order of $(1.8 \rightarrow 0.9)\times10^{-14}$ on time scales of 100 to 1,000 seconds. The neutral atmosphere is essentially non-dispersive at the AUSSAT radio-link frequencies and so it is absorbed into τ_1 and τ_2 of the phase transfer analysis. Changes in $\Delta\tau_A$ therefore affect the phase of fringes but do not affect the phase transfer system.

A4. <u>The design goals for the phase transfer system</u>

It has been shown in previous sections that the ionosphere can contribute delay rates of the order of 10^{-14} to both the astronomy signals and to the phase transfer system, and the neutral atmosphere can contribute rates of the same order to the astronomical signals. A delay rate of about 10^{-14} should be regarded as an upper limit to the permissible drifts on time scales greater than about 100 seconds introduced by instabilities in the electronics of the LBA and the phase transfer system.

Fluctuations in delay on short time scales are also important since they reduce the visibility of the fringes. The reduction factor is $\gamma = \exp[-\phi_\gamma^2/2]$ if Gaussian phase fluctuations with rms value $\phi\gamma$ are present during the integrations of the data. $\phi\gamma \leq 0.32$ radians is necessary to achieve $\gamma \geq 0.95$. The corresponding time jitter is $\Delta\tau_\gamma \leq 2.3$ pS rms at an observing frequency of 22.3 GHz.

Integration times of ≥ 10 seconds are envisaged for data from the LBA of the AT. These times are easily long enough to reduce the phase transfer error to below 2.3 pS but the equivalent local oscillator time fluctuations will be ≥ 10 pS if the local oscillators are generated from local rubidium standards. Figure 2 shows the way in which the phase transfer time errors and rubidium standard errors vary as a function of integration time. It is necessary for phase transfer measurements to be made, and corrections to the fringes to be applied, about once every second in order to keep the time fluctuations less than 2.3 pS.

Assuming that $\tau_D = 1$ second, then, from equation A1.5

$$\frac{1}{f_s}\left[\frac{C}{No}\right]^{-\frac{1}{2}} \leq 2.04\times10^{-11}$$

is required. The table shows the values of $\left(\frac{C}{No}\right)$ that are required and the rms time uncertainties $\sigma\tau$ corresponding to $\left(\frac{C}{No}\right)$ = 54 dBcHz for several values of f_s.

f_s (MHz)	$\left[\frac{C}{No}\right]$ (dBcHz)	$\sigma\tau$(pS) for $\frac{C}{No}$ =54dBcHz	No. of transponder combinations to achieve f_s
20	67.8	11.2	11
64	57.7	3.5	9
128	51.6	1.8	7
192	48.1	1.2	5
256	45.6	0.88	3
320	43.7	0.70	1

A5. Comment on the OTS-2 Results

The estimates of $\left(\frac{C}{No}\right)$ in the previous section seem low in in comparison with those quoted in AT/10.3/001 which were derived from the results of van Ardenne et al. (1983). There are two reasons for this. The first is that the results presented by van Ardenne et al. are basically rms deviations in the phase measurements (equivalent to $\{\langle(\Delta R_{xy}-\Delta R_{yx})^2\rangle\}^{1/2}$.
These deviations are twice as large as the corresponding uncertainties in the differences in phases of the frequency standards. The second reason is that the results are also a factor of two worse than theory predicts (see NFRA Note 341, A VIII-5; NFRA Note 395, Figure 1). Almost all of this loss can be attributed to image noise in PLL2 of their system, and to extra measurement noise due to insufficient resolution in the counter-timer used for the phase measurements. These losses can and should be avoided in an AUSSAT system.

A6. The frequency separation, f_s, of the pairs of tones.

The higher the frequency, f_s, the smaller is the value of $\left(\frac{C}{No}\right)$ needed to achieve a given value of $\sigma\tau$. However, f_s must be selected with regard to the restrictions imposed by AUSSAT. The nominal bandwidth of each channel on AUSSAT is 45 MHz and the channels have a centre frequency spacing of 64 MHz on a given polarisation. The channels using the other polarisation are offset in frequency by 32 MHz. The overlap of the band edges of cross-polarised channels may lead to carriers being restricted to the central 32 MHz of a channel in order to avoid cross interference. Values of f_s greater than 32 MHz are probably only achievable by putting the 2 tones of a pair through different channels.

The analysis in Appendix 1 shows no intrinsic loss in performance or in convenience in using different channels if the transponders are coherent ($\omega_A=\omega_B$) and shows the requirement to make twice as many phase measurements (i.e. including R_{xx} and R_{yy}) if they are not coherent. It would be sensible to use the same polarisation on both channels in order to simplify the ground terminals. The permissible values of f_s are thus

$$f_s = 64n \pm 32 \quad \text{MHz} \quad \text{where } n=0,1,\ldots 5 \text{ for}$$
one polarisation and n=0,1,...4 for the other.

Whilst a large value for n is desirable in order to minimise the $\left(\frac{C}{No}\right)$ that is required, the ability to fit into an allocation system for AUSSAT is reduced the large the value of n. The final column of the table in Section A.4 shows the number of transponder pairs that could be used to achieve a given value of f_s. The highest values of f_s (n=5) can only be achieved using two particular transponders. It would be safer and more flexible to choose n=3 or 4 (corresponding to f_s near 192 or 256 MHz). Any value of f_s grater than 0 and less than 352 MHz is possible but values near 64n MHz can be fitted into the channels in the greatest number of ways and would thus be easiest to accommodate if changes due to transponder failure were ever necessary. Values of f_s near 192 and 256 MHz seem to offer high performance and the flexibility to cope with any changes forced by equipment failure.

A.7. The stacking of pairs of tones

The pairs of tones from the different sites need to be stacked in a way which is compatible with the AUSSAT allocation system and which meets the performance targets of Appendix 4. It might seem necessary to choose non-commensurate values of the separations $(\omega_x - \omega_y)$ so that the intermodulation products generated by non-linearities in the transponder, do not interfere with any of the tones. Fortunately, since large numbers of intermodulation products are involved, it is shown in the next section that this is not necessary in order to meet the target specification. A simple system with equal offsets between adjacent pairs of tones can achieve the required performance.

A.7.1. Third-order intermodulatin products

The most troublesome non-linearities in narrowband systems are odd order

(i.e. with $V_{out} \propto \alpha V_{IN} + \beta V_{IN}^{k}$) where k = 3,5,7, etc.)

since these generate in-band and close in intermodulation products. The third order is usually most important. With an input containing several frequencies, f_1, f_1, f_3, etc a third-order non-linearity generates components such as $(f_1 - f_2)$ and $(f_1 + f_2 - f_3)$.

Consider a comb of N equally spaced frequencies, f_i each of the same aplitude, where

$$f_i = f_o + i\Delta f \quad \text{for } i=o,1,\ldots,(N-1) \text{ and } (N-1)\Delta f \ll f_o$$

A third-order non-linearity generates intermodulation products in the vicinity of f_0 at frequencies

$$f_j = f_0 + j\Delta f \quad \text{where } j = -N+1, -N+2, \ldots, 0, 1, \ldots, 2N-2$$

The number of ways of generating f_j is

$$n_j = (N+j)(N+j+1)/2 \quad \text{for} \quad -N+1 \leqslant j \leqslant -1 \text{ (below band)}$$

$$n_j = N(N+1)/2 + j(N-1) - j^2 \quad \text{for } 0 \leqslant j \leqslant N-1 \text{ (in band)}$$

$$n_j = (2N-j)(2N-j-1)/2 \quad \text{for} \quad N \leqslant j \leqslant 2N-2 \text{ (above band).}$$

The closest in "new" frequencies (f_{-1} and f_N) can be generated in

$$N_{-1} = n_N = N(N-1)/2 \text{ ways.}$$

The band-centre frequencies can be generated in

$$N_{(N-1)/2} = (3N^2+1)/4 \text{ ways if N is odd}$$

or $\quad N_{N/2} = N_{n/2} - 1 = 3N^2/4$ ways if N is even

If all the tones in the original spectrum are LdB below the 2 tone, third-order intercept point, then each intermodulation product will be 2LdB below the level of the tones. The total frequency spread of the products is $(3N-3)\Delta f$ = 3 times the spread of the tones.

A.7.2. <u>Timing errors due to intermodulation products</u>

The table in Appendix 4 shows that an overall $\frac{[C]}{[No]_T} \geqslant 48.1$ dBHz is required per tone for $\sigma\tau \leqslant 2.3$ pS with $f_s \geqslant 192$ MHz. The relative contributions of the uplink and the downlink to noise on the tone depend upon many factors including, for example, the transponder gain setting (NDG, page 24). Taking the worst possible transponder gain setting (high) and assuming that the up and down links degrade equally, then $\frac{[C]}{[No]_{U,D}} \geqslant 51.1$ dBHz are required.

A flux density of -90 $dBWM^{-2}$ is required to saturate a transponder in the high gain mode. Surface area of a sphere at satellite radius ≡ 162.4 dBM^2

saturation EIRP	= 72.4 dBW from the ground
Path attenuation	= -207 dB at 14.25 GHz
Atmospheric margin	= -1 dB
received power	= -135.6 dBW
AUSSAT receive G/T	= -3dB/K

$$\frac{[C]}{[T]_{TRANSP}} = 138.6 \text{ dB/K}$$

$$\frac{[C]}{[No]_{TRANSP}} = 90 \text{ dB Hz for the saturated transponder}$$

(K = -228.6 dB J/K).

The 2 tone, third-order intercept point is about 3 dB above the saturation level (NDG, page 27). Therefore, a single tone is L=93-51.2=41.9 dB below the intercept point and tone-to-tone intermodulation products are 2L=83.8 dB below the level of a tone.

For N=10 (=number of telescopes), there are 75 intermodulation products coincident with the central tones. Therefore, the (tone/net intermodulation product ratio is $83.8 - 10 \log_{10} 75 = 65.0$ dB assuming the products add incoherently. The resulting error in time transfer is, therefore,

$$\Delta\tau_{IM} = \frac{1}{2\pi f_s} 10^{-65.0/20}$$

$$= 0.45 \text{ pS rms for } f_s = 192 \text{ MHz}$$

This is far enough below the target of 2.3 pS that it is not necessary to enquire further about the coherency of the intermodulation products. Note, however, that there is not much scope for reducing σ_τ before intermodulation products become a nuisance.

A.7.3. Cross-modulation interference from other services sharing the transponder

This is potentially a serious threat to the viability of a phase transfer system. Fortunately, spectral dispersion of signals seems to be encouraged by the NDG. Scrambling of the digital signals and the use of high index FM with signal disposal spread the signal powers over the spectrum so that cross-modulation products just tend to raise the noise floor. If the signals are more or less evenly spread, this cross-modulation noise is a few dB below the "real" system noise and therefore not too important. Some signals such as those generated by voice activated circuits can appear essentially as unmodulated carriers and are more worrying. It is possible, however, to predict where these might arise and AUSSAT PTY can take some precautions to try to keep them at bay.

A.7.4. Net power/bandwidth requirements

Chapter 10 of Feher (1983) describes the subdivision of an INTELSAT transponder between many single-channel-per-carrier (SCPC) users. The numbers quoted for SCPC useage of AUSSAT in the Network Designers Guide (NDG) suggest that AUSSAT PTY has adopted the same specification. the NDG states that a single channel can transmit 56 Kbs^{-1} of data at error rates of $<3\text{x}10^{-9}$ with a system having

$$\left[\frac{C}{T}\right]_{SCPC} = -167.3 \text{ dbW/K}$$

$$\left[\frac{C}{N_o}\right]_{SCPC} = \left[\frac{C}{KT}\right]_{SCPC} = +61.3 \text{ dB Hz}$$

If the transponder is backed off 6 dB from saturation, then

$$\left[\frac{C}{No}\right]_{TRANS} \equiv 84 \text{ dB Hz}$$

and a transponder can support 200 channels. The transponder is power limited since only 9 MHz of the transponder bandwidth is occupied by channels spaced by 45 KHz.

A SCPC channel has sufficient power to carry at least 10 of the tones required for phase transfer. Since the offset ($f_x - f_y$) between the tone pairs can easily be reduced to a few KHz, 10 tones can also be squeezed into the 45 KHz channel allocation. A system using 2 SCPC channels separated by ≃192 or 256 MHz can transfer phase between up to 10 telescopes and use 0.5% of the useable power of each of two transponders.

The downlink requirements are as follows:

Transmitter EIRP		36 dBW saturated
Output back-off		-3 dB
Power dividing factor		-33 dB (1/2000 of useable power per tone)
EIRP/tone	=	0 dBW
Surface area of sphereal range		162.4 dBm^2
Gain/area conversion factor		43.4 dBm^{-2} ($4\pi/\lambda^2$ at λ = 2.4 cm)
Atmospheric margin		-1 dB
received power/antenna gain	=	-206.8 dBW per tone

But C/No ⩾ 51.1 dB Hz (or C/T ⩾ -177.5 dB W/K) is required

G/T ⩾ 29.3 dB is required

Assuming T = 300K ≡ 24.8 dB K, then

G ⩾ 54.1 dB is required.

This requires a 5 m diameter antenna on the ground at the operating wavelength of 2.4 cm.

The AT is unlikely to require phase to be transferred to 10 sites. It may be possible to negotiate the leasing of a fraction of each of 2 SCPC channels and hence reduce operating costs. Another possibility is to use the "spare" capacity for the transmission of control data. This is considered in more detail in Appendix 8.

A.7.5. Adjacent channel interference

If the tones are spaced evenly across the whole of a SCPC channel, then the outriding intermodulation products fall into the two adjacent channels. There are 165 outriders for N=10 in each of the immediately adjacent channels. The net power spilling over is

$$+ 93 - 3L + 10 \log_{10}(165) = -10.5 \text{ dBHz}$$

compared to the 61.3 dBHz legitimately available for the channel. The interference level is thus down by 71.8 dB with respect to the signal and can therefore be completely ignored.

A.8. The transmission of control and monitoring data via AUSSAT

A SCPC channel on AUSSAT can carry 56 Kbs_{-1} of data with an error rate $\leqslant 3\times10^{-9}$ by using 4 PSK modulation with data coding and scrambling. It is possible, in principle, to subdivide the power and bandwidth allocated to such a channel into several lower capacity channels which could also be accessed independently. A division of both by a factor of 10 would leave sufficient carrier power per subchannel to support either a phase transfer tone as outlined in Appendix 7, or to enable data to be transmitted at a rate up to 4.8 Kbs_{-1}.

Two SCPC channels are required to allow the tones of a pair to be separated by 192 or 256 MHz. There is enough capacity in these two channels to support 6 telescopes with 2 tones and a data channel each. At 200 SCPC channels per transponder and \$A2 M $year^{-1}$, the running costs for a phase and control data transfer system would be approximately \$A20 K $year^{-1}$. It is not possible to use the carrier recovered from the data demodulation process for phase transfer since the jitter on the recovered carrier caused by intersymbol interference is too large.

A.9. AUSSAT stability requirements

It has been established that short term fluctuations in $\Delta(\phi_s^x - \phi_s^y)/2\pi f_s$ should be less than 2.3pS and that the rate of change should be less than 10^{-14} over time scales of about 100 secs. The corresponding limits on the phase shifts in AUSSAT are

$$(\langle(\Delta\Phi_{AUSSAT} - \langle\Delta\Phi_{AUSSAT}\rangle)^2\rangle)^{½} \ll 5.5\times10^{-3} \text{ radians}$$

and $\Delta\dot{\Phi}_{AUSSAT} \ll 2.4\times10^{-5}$ radians sec^{-1} assuming f_s = 192 MHz.

If all the tones can be stacked so that $(\omega_x - \omega_y)/2\pi \leqslant 45$ kHz, then

$$\Delta\Phi_{AUSSAT}/(\omega_x - \omega_y) = \left[\frac{\delta\Phi_u}{\delta\omega}^b (\omega_y + \omega_s) - \frac{\delta\Phi_u}{\delta\omega}^A (\omega_y) + \frac{\delta\phi_D}{\delta\omega}^B (\omega_y + \omega_s - \omega_b) - \frac{\delta\Phi_D}{\delta\omega}^A (\omega_y - \omega_A)\right]$$

has fluctuation limits << 20 nS, and rate limits << 9×10^{-11}

The first two terms inside the square brackets are the group delays of the uplink channels. The major contributors to these delays are the filters which define the bandpasses. Near the band centre of a filter, the group delay is approximately $\simeq 1/\pi \times$BANDWIDTH = 7 nS for a bandwidth of 45 MHz. Fluctuations of these delays are obviously not a problem. The difference in the rates of change of delay should be less than 1.2% per sec which should not a problem either, especially since some degree of cancellation might be expected.

The third and fourth terms inside the square brackets are the group delays of the downlinks. Filters contribute here but the same comment applies as above. The travelling wave tube amplifiers (TWTA) contribute significant extra delay. Guessing a tube length of $\leqslant$30 cm and a slow wave velocity of $\geqslant$0.05 C ($\simeq$ 1 Kev electrons), gives a tube delay of $\leqslant$20 nS. Fluctuations in this should not be a problem. The differential rate limit is 0.45% per second corresponding to a voltage rail limit of 0.9% per second. This again should not be a problem provided that either common rails are used or else the voltages are stabilized.

Another problem that might arise with the TWTA is caused by AM to PM converson in the tube, i.e. changes in drive level affecting the phase shift through the device. The "coefficient" is of the order of degrees per dB (4.6° dB^{-1} for AUSSAT). However, all the tones through a transponder are likely to be similarly affected and hence these shifts should cancel out to a large extent. The phase shifts would cause temporary glitches in the phase measurement but that should not cause any problems since the shifts are likely to be $\ll \pi$ radians. The unit of drive change is 1/200 of the useable power i.e. only 0.02 dB so the units of phase change are very small in any case.

A.10. <u>Ground station stability requirements</u>

The ground stations are not specifically included in the analysis in Appendix 1. It is obvious that extra phase shifts

$$\Delta\phi_x \text{ at } \omega_x \text{ and } \Delta\phi_{xs} \text{ at } \omega_x + \omega_s$$

result in extra contributions to R_{xy} amounting to $\Delta\phi_{xs} - \Delta\phi_x$

The affect on phase transfer is then

$$\Delta(\phi_s{}^x - \phi_s{}^y) = (\Delta\phi_{xs} - \Delta\phi_x)/2$$

Thus, the differential phase shift limits are

$\ll 5.5\times10^{-3}$ radians for fluctuations
and $\ll 2.4\times10^{-5}$ rad sec^{-1} for rates

For dispersionless delays such as in a TWTA, the group delay limits are

$$\ll 4.6 \text{ pS} \equiv 0.023\% \text{ for fluctuations}$$
$$\text{and } \ll 2\text{x}10^{-14} \equiv 0.0001\% \text{ sec}^{-1} \text{ for rates.}$$

Both of these might be difficult to meet. It would be better to avoid using a TWTA. The delay through a GaAs FET amplifier is probably 2 orders of magnitude less than through a TWTA and should be able to meet these specifications.

If it is assumed that the transmitter and receiver bandwidths are at least 256 MHz, then the group delay stability requirements are

$$\ll 0.37\% \text{ for fluctuations}$$
$$\ll 0.002\% \text{ sec}^{-1} \text{ for rates}$$

Neither of these two should be too difficult to satisfy provided that the equipment is housed in a reasonably draught-proof and temperature-stable environment. Somce investigation of antenna mounted units such as those offered by NEC may be necessary.

There is, however, an additional complication of the transmitters. The NDG (page 26, Table 2.5) seems to require the transmitter response to roll off by at least 19 dB at ±41.5 MHz and by at least 25 dB at ± 64 MHz relative to the transponder band centre. In this application, the ground station will be working two transponders and it will be necessary to have staggered band-pass filters in the transmitter outputs. If each of these bands is 45 MHz wide, the approximate "Q" of the filters is about 320. The rate of change of phase shift through a filter is

$$\frac{\delta\phi}{\delta\tau} = \frac{\delta[2\pi f\tau]}{\delta\tau}$$

$$= 2Q\left[\frac{\delta f}{f\delta\tau}\right] \text{ since } \tau \equiv \frac{1}{\pi B} = \frac{Q}{\pi f}$$

The rate limit of $2.4\text{x}10^{-5}$ rad s^{-1} translates into a limit on the differential rate of change of tuning of the filters of

$$\left[\frac{\delta f}{f\delta\tau}\right]_1 - \left[\frac{\delta f}{f\delta\tau}\right]_2 \ll \frac{1}{2Q}\frac{\delta\phi}{\delta\tau} = 3.8\text{x}10^{-8} \text{ sec}^{-1}$$

The most likely cause of drift is temperature variations. The requirement is

$$\left\{\left[\frac{\delta f}{f\delta T}\right]_1 - \left[\frac{\delta f}{f\delta T}\right]_2\right\} \frac{\delta T}{\delta \tau} << 3.8\times10^{-8}\ \text{sec}^{-1}$$

if good thermal contact is maintained between the filters. Guessing that $\left[\frac{\delta f}{f\delta T}\right] \equiv 3.8\times10^{-5}$ per °C is achievable and that two similarly designed and constructed filters track one another to the 20% level, then $\frac{\delta T}{\delta \tau} << 5\times10^{-3}\ \text{K sec}^{-1}$ is required. Good thermal contact between and thermal lagging around, the filters are going to be necessary to achieve that sort of performance without resorting to active and accurate temperature control.

The output power requirements of the transmitter are quite moderate. An EIRP of 33.4 to 43.4 dBW per tone (depending on the transponder gain setting) is required to grab 1/2000 of the backed-off transponder power. Using a 5 m diameter dish with a gain of 54.3 dB gives a transmitter power of -20.9 to -10.9 dBW per tone or -16.1 to -6.1 dBW for 2 tones and a data channel. The NDG (page 34) requires spurious outputs from the transmitter to be $\leq$-60 dBW in any 4 KHz band when measured at the antenna interface. This is -39.1 dB with respect to the level of a tone. The two-tone, third-order intercept point of the GaAs FET amplifier should therefore be > -10.9+(39.1)/2 = 8.7 dBW in order to satisfy this criterion. Since the 1 dB gain compression point is usually about 10 dB below the intercept point, an amplifier with a 1 dB compression point at about -0 dBW is required. (This is probably being too conservative. See the NDG Section 3.11, page 35).

The requirements on the phase measuring equipment are discussed in Appendix 11. The only other requirement of the ground station receivers apart from those already discussed is the need for linearity. The receivers must not generate intermodulation products at greater relative levels than those generated in the transponders on AUSSAT. The intercept point must be at least 42 dB above the level of a single tone.

A.11. Tone-separation-and-measurement

A measurement system based on the one used by van Ardenne et al. (1983) is envisaged but with a few minor differences. It is assumed that the tone pairs are offset at intervals of 4 KHz. Figure 3 is a schematic diagram of a system to separate and process the tones prior to the measurement of phase. The intention is to treat both members of a pair of tones identically as far as possible in order to take advantage of some compensation of drifts, and to avoid the use of high Q filters. Image reject filters are used for the conversions down to 5 KHz to prevent degration of signal-to-noise ratios.

Tentative signal levels are marked on the diagram on the assumption that the mixers that are used have a two-tone, third-order intercept point at $\geqslant$+10dBm. The level of a tone must always be at least 42 dB below the relevant intercept point where ever cross modulation between tones is possible.

The bandwidth is limited to 4 KHz in order to get a S/N $\geqslant$12 dB at the phase detector of the PLL. The filters also reject adjacent tones and tone images. Assuming delays τ_1 and τ_2 through these 5 KHz filters, then the differential phase error between tones of a pair is

$$\phi = 2\pi f_m(\tau_1 - \tau_2), \quad f_m = 5 \text{ KHz}$$

$$\delta\phi = 2\pi f_m \Delta\tau\left[\frac{\delta f_m}{f_m} + \frac{\delta(\Delta\tau)}{\Delta\tau}\right] \text{ where } \Delta\tau = \tau_1 - \tau_2$$

$$<< 10^{-3} \text{ rads for } \Delta\tau < 0.8 \text{ pS at 192 MHz}$$

$$\tau_1 = \tau_2 \simeq 1/\pi B = 80 \ \mu S \text{ for 4 KHz bandwidth}$$

$$\delta(\Delta\tau) = 3\text{x}10^{-8} \text{ sec}$$

$$\text{or } \frac{\delta(\Delta\tau)}{\tau_1} = 4\text{x}10^{-4}$$

Temperature coefficients of $\leqslant 10^{-4} k^{-1}$ are attainable and matching the TC's to $\simeq$ 20% is likely. With care, the narrow band filtering should not cause too much drift. Note however that $\delta f_m \leqslant 2$ Hz is also required.

Uncertainties in frequency of up to ± 20 KHz exist due to drift in the translation oscillator on AUSSAT. This offset is common to all the tones and is removed by appropriate tuning of the VCXO in the down converter from the first IF at 1 GHz. This VCXO replaces the VCXO in the phase measurement module which tracks the pair of tones from the master station. By this means $\delta f_m \simeq 0$ is achieved.

An intelligent controller is required to lock the system onto the correct tones. The controller tunes the master VCXO until data is decoded successfully. A byte or string in the date-message identifies the frequency of the channel to the controller which adjusts the tuning if it is not the expected channel. Once the controller has got the tuning correct, it allows the "reference" loop to close. The reference loop needs to be a few hundred hertz wide, the signal loop only a few hertz wide. The VCXOs in the measurement modules should have a tuning range of a few hundred hertz to prevent locking onto the "wrong" tones. The controller is also required to multiplex the time/interval meter ($\leqslant$ 0.1 μS resolution) between the various tones, to read out the TIM, to adjust appropriately the samples which have over/under ranged, and to average to the 1 second level.