Carrier Phase Ambiguity Resolution in GNSS-2

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BIOGRAPHIES

Börje Forssell is Professor of Navigation at the Norwegian University of Science and Technology in Trondheim, Norway. In 1996-97 he has been working as a consultant at ESA/ESTEC, on leave from the University. His research activities at ESTEC were focussed on GNSS-2.

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Bob Harris has been with ESA since 1970 and was responsible for transmission techniques in ESA's Directorate of Telecommunications from 1978 to 1997. He worked for two years on GNSS-2.

ABSTRACT

The accuracy requirements of future satellite-based positioning applications are expected to necessitate the use of carrier-phase techniques. These must at the same time be rapid and robust. The signal designs of GPS and GLONASS limit what can be achieved, but it may be possible to improve the situation significantly in the next generation of systems. As part of its contribution to worldwide studies of GNSS-2, ESA has been studying a modified signal design using three suitably spaced frequencies. This paper describes the TCAR method, presents a preliminary analysis of its performance and describes a planned programme of experiments to validate it.

The TCAR method is simply an extension of the well known *widelaning* technique. Three carriers are spaced such that the frequency differences between them form a steady progression, in four steps, from the chip rate to one of the carrier frequencies. The two "outer" carriers form a conventional *widelane,* while the closest two form a *super-widelane.* As long as the noise and bias differences between steps are sufficiently small, the integer wavelength ambiguity inherent in the carrier-phase pseudorange estimate at each step can be solved using the pseudorange estimate from the previous step. In principle this can be carried right down to the carrier wavelength, but residual biases become increasingly difficult to treat in later steps and differential or estimation and compensation techniques are needed.

1. INTRODUCTION AND SCOPE

The last thirty years have seen an enormous increase in the interest in position determination and navigation — and in the accuracies achieved, principally due to the use of satellite-based techniques developed for military purposes. The opening up of the GPS and GLONASS systems to civilian users has led to an explosive growth in the range of applications and to significantly increased performance (accuracy, speed of solution, robustness and reliability) requirements, which go well beyond the original targets.

Although the far-sighted designs of the existing systems, coupled with advances in technology and signal processing techniques have allowed dramatically improved performances to be achieved, the constraints imposed by the design epoch and the military applications make further improvements increasingly difficult and costly. It is however expected that civilian demands will continue to increase and there are moves to develop a civilian-controlled Global Navigation Satellite System (GNSS).

First steps in this direction are already being taken with the implementation in the USA, Europe and Japan, of systems based on GPS and GLONASS, using networks of earth-stations to provide improved system integrity. These systems are largely oriented towards the requirements of civil aviation. In the slightly longer-term however, there is pressure to develop a second-generation GNSS which should be fully civilian-controlled and offer significantly improved performance for a wide range of applications.

Although the institutional framework for such a GNSS-2 system is far from clear, studies are under way to identify suitable system characteristics. As part of this work, The European Space Agency (ESA) is sponsoring a wide-ranging set of studies and experiments. The start date assumed for GNSS-2 is the second decade of the new millenium, and the studies do not assume that the design of the new system will necessarily be tied to those of the present GPS or GLONASS.

Many of the required accuracy improvements can only be satisfied by the use of carrier-phase techniques and we expect that their use will become routine. At the same time, requirements on speed and reliability of solutions will continue to increase. One of the topics being studied in the ESA programme is a modified signal design to make it easier and faster to resolve the integer wavelength ambiguity inherent in such measurements.

The studies are at an early stage and this paper should be seen as a progress report. Section 2 briefly reviews the current situation. Section 3 introduces the basic principles underlying the new structure and sections 4 and 5 provide observation equations and noise performance analyses. Section 6 outlines plans for in-orbit experiments. It is stressed that this paper focuses on ambiguity-resolution (AR) rather than on the absolute accuracy of the range measurements.

2. CURRENT METHODS

GPS and GLONASS ranging is based on the use of two carriers and two codes with different lengths and chip rates. Determination of the code phase allows a rapid and unambiguous, but relatively coarse, estimate of the range. Measurement of the carrier phase allows the range to be resolved finely, but with an uncertainty concerning the number of wavelengths involved.

The large ratio ("gap") between the spreading-code chip rate and the carrier frequencies means that the carrier-phase ambiguity can only be solved by statistical methods, involving complicated and time-consuming searches in multi-dimensional space. Considerable efforts and resources have been — and are still being — spent on the development and refinement of sophisticated and reliable methods to solve these problems under different conditions.

The integer ambiguity is usually resolved by first defining a search volume by means of the code and then trying possible values of integer combinations against some validation criteria. In order to make the search volume as small as possible, the higher chip rate should be used, but this code is usually inaccessible to civilian users. Also *widelaning*, i.e. phase differencing between the carriers, can be used to reduce the search volume, but this is hampered by the same restrictions, leading to increased noise levels.

In any case, the solutions inevitably take some time to find and to validate. Validation criteria are based on certain differences between false and correct solutions, which may require measurement epochs of the length of minutes to give acceptable confidence levels. This also increases the probability of cycle slips.

The on-going discussion regarding the possible addition of a third navigation frequency (sometimes called L5) in future GPS satellites has also touched upon the phase-ambiguity problem. A proposal in this respect, very much in line with the method described below, is presented in [1]. That method however starts from the basis of frequencies chosen for other purposes and assesses the performance achievable. Here we try to identify the criteria for optimising the choice of frequencies.

3. PRINCIPLE OF THE "GAP-BRIDGING" THREE-CARRIER METHOD

As indicated above, the main obstacle to fast and reliable carrier-phase ambiguity resolution is the large *wavelength gap* between code chip and carrier. The widelaning technique, sometimes used in GPS and other systems, goes some way towards bridging this gap, by using the difference between the two carrier frequencies. The large frequency spacing required for good ionospheric delay compensation still however leaves a large gap between the "widelane" difference-frequency wavelength and that of the chip.

One obvious way to reduce the size of the gap would be to increase the chip rate and hence the signal bandwidth. As well as easing the ambiguity-resolution problem this would give better code-phase resolution. It would however give rise to a number of technical and regulatory problems and a "lower chip rate" solution appears to be preferable.

The purpose of the three-carrier (TCAR) method described here is to make reliable real-time AR feasible, by providing a set of overlapping steps from a relatively modest code chip rate to the carrier frequency. The underlying principle is simply an extension of the widelaning technique.

Fig. 1 illustrates some of the relationships involved using three carriers. The two "outer" carrier frequencies (f_1 and f_3) are spaced by f_{13}, which is fairly small compared with either frequency. The "third" frequency (f_2) is separated from f_1 by f_{12}, which is in between the chip rate and f_{13}. The wavelengths of the two difference frequencies thus provide intermediate steps between the chip and carrier wavelengths.

The "outer" frequencies, f_1 and f_3, can be compared to the GPS L1 and L2 carriers, with f_{13} corresponding to the "widelane". By extension, f_{12} can be thought of as a "super widelane".

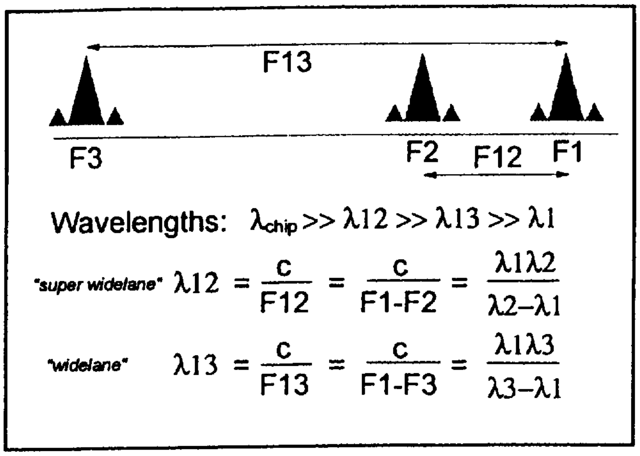

Fig. 1: The Wavelength-Gap-Bridging Method

The selection of the three carrier frequencies depends on the ratio between f_1 and the chip rate and the "obvious" intermediate steps form a geometric progression between them. The choices will however usually be influenced by other factors, such as the (non-) availability of suitable frequencies or a wish, for instance, to keep open the possibility of dual-frequency ionospheric-delay correction, even though the corresponding frequency separation f_{13} may be larger than optimum for AR purposes. To some extent, "non-optimum" frequency spacings can be compensated by changing the balance between the powers of the carriers.

Fig. 2 shows the basic principles of the TCAR method. In Step 1, the pseudorange is estimated from the code phase, using the carrier at f_1. This can be considered to consist of a "reference quantity" (REF), which is a biased version of the true geometric range, together with "noise terms" which include other biases. The biases in REF propagate through all steps in the process. They have to be removed, or at least reduced later, in order to achieve the desired accuracy. In some cases, as shown in the next section, it will also be necessary to reduce them for the TCAR method to function properly.

In Step 2, carrier-phase measurements are made at f_1 and f_2, using a common reference oscillator. By taking the difference between the two phases, each expressed in terms of cycles at its own carrier frequency, we can form a carrier-phase pseudorange estimate in terms of the wavelength λ_{12} at the super-widelane difference frequency f_{12}, except that the number N_{12} of complete wavelengths is not known.

The pseudorange estimate in Step 2 can be shown to have a bias which is very close to REF. Provided the measurement noises at the two steps and differences between the two biases are small compared to λ_{12}, the

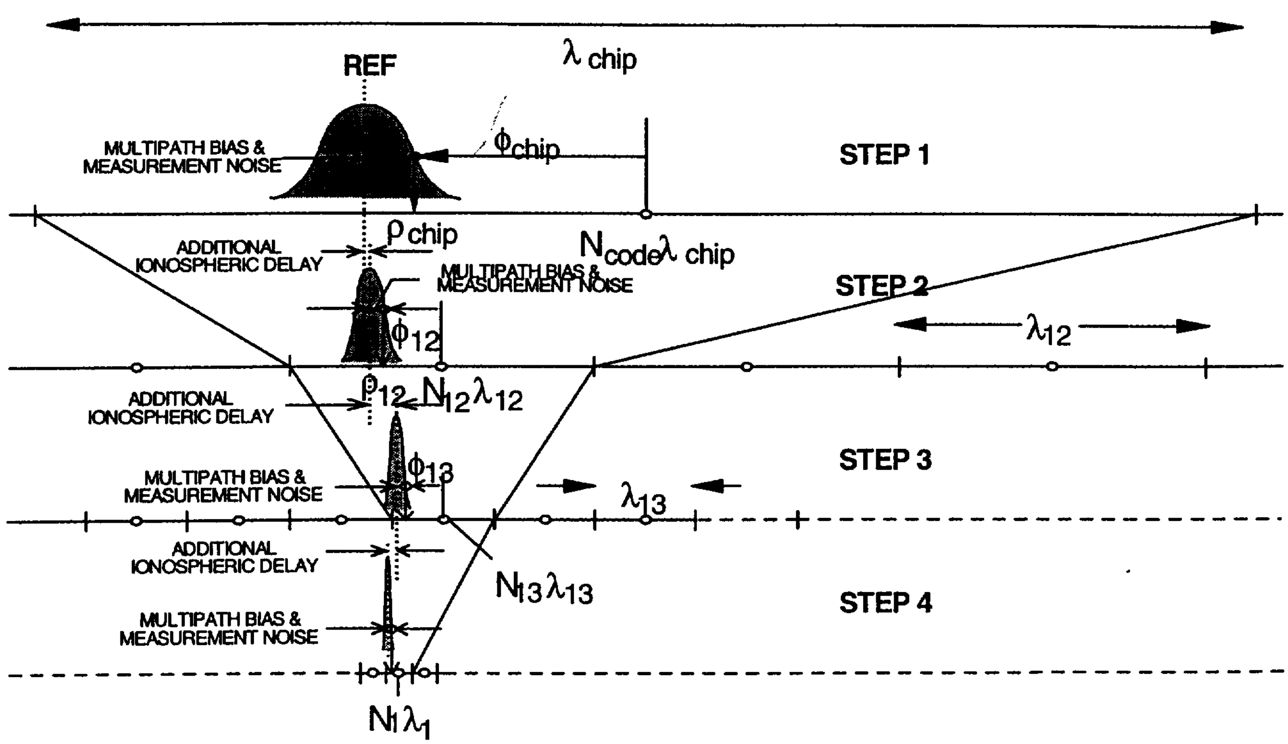

Fig. 2: Steps in the Successive Refinement Process

Step 1 and Step 2 pseudorange estimates can be combined and N_{12} estimated with good reliability. Exactly how small the noise and bias differences need to be depends on "confidence" (reliability) requirements and on the balance between noise and biases.

Step 3 is very similar to Step 2 except that now the starting point is the Step 2 estimate, which has a much smaller measurement noise than that at Step 1. An additional carrier phase measurement is made at f_3 and subtracted from that at f_1 to form a new ambiguous pseudorange estimate in terms of λ_{13}. The Step 2 and Step 3 estimates are then combined to estimate the number of wavelengths at f_{13}.
Finally, in Step 4, the carrier phase measurement at f_1 is related to the Step 3 estimate to estimate the number of cycles at f_1.

It should be carefully kept in mind that resolution and accuracy are two different things. Thus, the "gap-bridging" described can be successfully carried out without necessarily achieving final accuracies in the order of fractions of λ_1 (i.e. if the bias or noise errors at any step are too large, the subsequent step(s) in Fig. 2 may be irrelevant.).

4. THE OBSERVATION EQUATIONS

The basic principles of the TCAR method are illustrated next through simplified observation equations. The signals are assumed to be transmitted coherently and the receiver is assumed to perform coherent operations across its frequency channels.

STEP 1:

The observed pseudorange ρ for frequency f_1 and a particular GNSS satellite is given by:

$$\rho = r - \left(d_{clk_{sat}} - d_{clk_{rx}}\right) + d_{tropo} + \frac{K}{f_1^2} + d_{mp_{code}} + n_{code} \qquad (1)$$

where:

- r is the true geometric range,
- $d_{clk_{sat}}$ is the path length corresponding to the satellite clock error relative to true GPS time,
- $d_{clk_{rx}}$ is the path length corresponding to the receiver clock error relative to true GPS time,
- d_{tropo} is the excess path length due to tropospheric delay,
- K is the ionospheric delay constant, directly proportional to the Total Electron Content along the path,
- $d_{mp_{code}}$ is the code-phase multipath range error

and

- n_{code} is the measurement noise on the code phase observations, expressed in metres.

The first five terms in (1) form the common "reference quantity" REF, introduced in the previous section and so:

$$\rho = REF + \left(d_{mp_{code}} + n_{code}\right) \qquad (2)$$

The bias terms grouped together in REF are all suitable for correction by one means or another. Normally it will be preferable to do this at as late a step as possible, to minimise the noise on the final estimate.

As long as the bias terms in (1) and the user's position uncertainty are small compared with the wavelength of the code structure (usually many chips), the pseudorange in (1) can be determined unambiguously. In practice this will normally only require solving for the receiver clock error, by using a redundant satellite measurement.

Multipath errors are frequency-, site- and path-dependent. They are normally uncorrelated and so cannot be reduced by differential methods. They must be kept small by careful site selection (when possible), appropriate receiving antenna design and proper signal processing in the receiver. They are likely to set limits on the accuracy ultimately achievable.

STEP 2:

First, the phases of the received carriers at f_1 and f_2 are measured using a common reference. As the satellite generates the carriers coherently with the code structure, the corresponding phase of each transmitted carrier can be estimated using the receiver clock, and the difference between the transmitted and received phases can be used to compute the carrier-phase pseudorange:

$$\lambda_1(N_1 + \phi_1) = r - d_{clk_{sat}} + d_{clk_{rx}} + d_{tropo} - \frac{K}{f_1^2} + d_{mp_1} + n_1 \qquad (3)$$

where:

- $\lambda_1 = c/f_1$ is the carrier wavelength,
- N_1 is the integer ambiguity (unknown number of wavelengths) at f_1,
- ϕ_1 is the difference between the carrier phase measured at the receiver and that computed at the transmitter,
- d_{mp_1} is the carrier-phase multipath delay error at f_1

and

n_1 is the carrier-phase measurement noise error.

The negative sign in the ionospheric delay term represents a phase advance.

As already explained, in TCAR, a "super-widelane" observable is formed from the phase measurements at f_1 and f_2. The observation equation for the super-widelane is very similar to that of the carrier phase:

$$\lambda_{12}(N_{12}+\phi_{12}) = r - d_{clk_{sat}} + d_{clk_{rx}} + d_{tropo} + \frac{K}{f_1 f_2} + d'_{mp_{12}} + n'_{12} \qquad (4)$$

where:

$N_{12} = N_1 - N_2$,

$\phi_{12} = \phi_1 - \phi_2$,

λ_{12} is the super-widelane wavelength, defined by:

$$\frac{1}{\lambda_{12}} \equiv \frac{1}{\lambda_1} - \frac{1}{\lambda_2} \qquad (5)$$

and

$d'_{mp_{12}}$ and n'_{12} are respectively the differences between the multipath and noise terms at f_1 and f_2, each scaled up to λ_{12}.

The change in the sign of the ionospheric term from (3) to (4) results from the fact that each phase shift is expressed in cycles at its own carrier frequency. It corresponds to a relative delay between the phases of the carriers [2].

We can now solve the integer ambiguity N_{12} associated with the super-widelane phase-difference estimate φ_{12}, using the previous code range estimate ρ.

Combining (1) and (4) gives:

$$\lambda_{12}(N_{12}+\phi_{12}) = \rho + \frac{K}{f_1^2}\left(\frac{f_1}{f_2}-1\right) + \left(d'_{mp_{12}} - d_{mp_{code}}\right) + \left(n'_{12} - n_{code}\right) \qquad (6)$$

The estimate of the super-widelane ambiguity is thus:

$$\overline{N}_{12} = round\left[\frac{1}{\lambda_{12}}\rho - \phi_{12}\right] \qquad (7)$$

where:

$round[x]$ indicates the nearest integer to x.

Eq. (6) includes an additional ionospheric delay bias, relative to REF, because of the frequency difference of:

$$\frac{1}{\lambda_{12}}\frac{K}{f_1^2}\left(\frac{f_1}{f_2}-1\right) \qquad (8)$$

and a "measurement noise" of:

$$dN_{12} = \frac{1}{\lambda_{12}}\left[\left(d'_{mp_{12}} - d_{mp_{code}}\right) + \left(n'_{12} - n_{code}\right)\right] \qquad (9)$$

where both (8) and (9) are expressed as fractions of λ_{12}.

For successful ambiguity resolution, the sum of (8) and (9) must be less than 0.5.

Satellite and receiver clock errors and tropospheric delay are contained in REF and do not affect the AR process at all. As f_1 and f_2 are very close, the additional ionospheric delay bias, (8) is very significantly attenuated and even "raw" measurements should not give problems for the Step 2 AR process.

The thermal noise will normally be dominated by n_{code}, but this will usually be small compared with λ_{12}. The critical "unknowns" are likely to be the multipath components.

The Step 2 estimate (4) can now be re-written as:

$$\lambda_{12}\left(\overline{N}_{12}+\phi_{12}\right) = REF + \frac{K}{f_1^2}\left(\frac{f_1}{f_2}-1\right) + d'_{mp_{12}} + n'_{12} \qquad (10)$$

which serves as the reference for Step 3.

<u>STEP 3</u>:

This is similar to Step 2, but involves a carrier phase measurement at f_3, rather than f_2 and the formation of the corresponding widelane quantities.

The Step 3 expression equivalent to (6) is:

$$\lambda_{13}(N_{13}+\phi_{13}) = \lambda_{12}(\overline{N}_{12}+\phi_{12}) + \frac{K}{f_1^2}\left(\frac{f_1}{f_3}-\frac{f_1}{f_2}\right) + \left(d'_{mp_{13}} - d'_{mp_{12}}\right) + \left(n'_{13} - n'_{12}\right) \qquad (11)$$

The estimate of the widelane ambiguity is:

$$\overline{N}_{13} = {}_{round}\left[\frac{1}{\lambda_{13}}\lambda_{12}(\overline{N}_{12} + \phi_{12}) - \phi_{13}\right] \quad (12)$$

Eq. (11) includes an ionospheric bias relative to the Step 2 estimate, of:

$$\frac{1}{\lambda_{13}}\frac{K}{f_1^2}\left(\frac{f_1}{f_3} - \frac{f_1}{f_2}\right) \quad (13)$$

and a "measurement noise" of:

$$dN_{13} = \frac{1}{\lambda_{13}}\left[\left(d'_{mp_{\phi13}} - d'_{mp_{\phi12}}\right) + \left(n'_{13} - n'_{12}\right)\right] \quad (14)$$

The sum of (13) and (14) must be less than 0.5 for successful AR. As f_1 and f_3 are much more widely separated than f_1 and f_2, Step 3 is clearly more sensitive to ionospheric effects than Step 2. For typical L-band frequencies, the nett ionospheric delay error, K/f_1^2, would need to be less than about 1 metre. In most circumstances some form of correction will be required. At this step, differential measurement will often be appropriate, but an alternative method, described in Step 4, may also be applicable.

Again, thermal noise is not likely to be a problem, but multipath needs to be kept small.

The Step 3 estimate can be written as:

$$\lambda_{13}\left(\overline{N}_{13} + \phi_{13}\right) = REF + \frac{K}{f_1^2}\left(\frac{f_1}{f_3} - 1\right) + d'_{mp_{13}} + n'_{13} \quad (15)$$

which serves as a reference for Step 4.

STEP 4:

In this step, the carrier-phase, measured at f_1, is used on its own, and the corresponding pseudorange is given by (3). Combining this with (15) we have:

$$\begin{aligned}&\lambda_1(N_1 + \phi_1)\\ &= \lambda_{13}(N_{13} + \phi_{13}) - \frac{K}{f_1^2}\left(1 + \frac{f_1}{f_3}\right) \quad (16)\\ &+ \left(d_{mp_1} - d'_{mp_{13}}\right) + \left(n_1 - n'_{13}\right)\end{aligned}$$

Similarly to (12) - (14), N_1 can be estimated correctly, from (16), provided the sum of the "extra" ionospheric bias, multipath and measurement noise terms is less than $\lambda_1/2$.

Unlike Steps 2 and 3, the ionospheric delay terms in (3) and (15) have opposite sign. The magnitude of the "raw" (ionospheric delay) bias jump between Steps 3 and 4 is therefore more than twice that of the ionospheric delay bias affecting individual code-phase or carrier-phase pseudoranges. As λ_1 is very small, this is potentially a serious problem. For typical L-band frequencies the nett ionospheric delay error K/f_1^2 would need to be reduced to a few cm.

Because ionospheric delay is path-dependent, differential measurements can only achieve the required reduction with very short baselines. In other cases, accurate estimation of K may be necessary.

One possibility may be to replace N_{13} in (11) by the estimate from (12) and solve for K, giving:

$$\frac{K}{f_1^2} = \frac{\begin{bmatrix}\lambda_{13}\left(\overline{N}_{13} + \phi_{13}\right) - \lambda_{12}\left(\overline{N}_{12} + \phi_{12}\right)\\ -\left(d'_{mp_{13}} - d'_{mp_{12}}\right) - \left(n'_{13} - n'_{12}\right)\end{bmatrix}}{\left(\frac{f_1}{f_3} - \frac{f_1}{f_2}\right)} \quad (17)$$

The first two terms in the numerator are formed from measured and previously estimated quantities. The multipath and noise terms are unknown, but include elements from Step 2, which are likely to be large relative to the required accuracy. If the mean value of these terms is sufficiently small, the noise can be reduced by making a fit to previous measurements. K can then be estimated as:

$$\frac{\overline{K}}{f_1^2} = \frac{\left(\overline{\lambda_{13}\left(N_{13} + \phi_{13}\right) - \lambda_{12}\left(N_{12} - \phi_{12}\right)}\right)}{\left(\frac{f_1}{f_3} - \frac{f_1}{f_2}\right)} \quad (18)$$

Assuming that the ionospheric bias has been suitably reduced, the carrier ambiguity can be solved as:

$$\overline{N}_1 = \frac{1}{\lambda_1}\lambda_{13}\left(N_{13} + \phi_{13}\right) - \phi_1 \quad (19)$$

with an error given by:

$$dN_1 = \frac{1}{\lambda_1}\left(\begin{array}{l}\frac{-d\bar{K}}{f_1^2}\left(1+\frac{f_1}{f_3}\right) \\ +\left(d_{mp_1} - d'_{mp_{13}}\right)+\left(n_1 - n'_{13}\right)\end{array}\right) \quad (20)$$

which must be smaller than 0.5 for a final and complete TCAR resolution.

5. NOISE ANALYSIS

As indicated above, the noise and bias terms at each step should be sufficiently small that the probability of making an error in the estimate of the integer N at the following step is negligibly small. Ignoring the residual bias for the moment, we need only consider the distribution of the noise. We assume a noise standard deviation, at each step, of $1/2\nu$ of the wavelength at the following step.

Thus:

$$\sigma_k \leq \frac{\lambda_{k+1}}{2\nu} = \frac{\lambda_k}{2\nu R_k} \quad (21)$$

where:

$R_k = \frac{\lambda_k}{\lambda_{k+1}}$ is the wavelength ratio.

The noise standard deviation (in metres) for code-phase pseudorange measurements is:

$$\sigma_{code}^2 \approx \frac{B_{code}\lambda_{code}^2}{2[C/N_0]_1} \quad (22)$$

where B_{code} is the noise bandwidth of the code tracking loop and $[C/N_0]_1$ is the carrier noise density on carrier 1.

Consequently, the required signal-to-noise-density ratio on carrier 1 to meet the requirement of (21) is given by:

$$\left[\frac{C}{N_0}\right]_1 = 2\nu^2 B_{code} R_1^2 \quad (23)$$

Similarly to (22) for a single carrier-phase measurement:

$$\sigma_{carrier}^2 \approx \frac{B_{carrier}\lambda_{carrier}^2}{4\pi^2[C/N_0]_\phi} \quad (24)$$

Substituting from (21):

$$\left[\frac{C}{N_0}\right]_\phi = \frac{\nu^2 B_{carrier} R_k^2}{\pi^2} \quad (25)$$

Eq. (25) applies to single frequency measurements. Wide-laning increases the phase noise as:

$$n_{ki}^2 = n_k^2 + n_i^2 \quad (26)$$

where:

n_k and n_i are the phase noises at f_k and f_i respectively and

n_{ki} is the wide-lane noise,

all expressed in cycles at the respective frequency. Taking single or double differences also increases the measurement noise and the required C/N_0.

Suitably modified versions of (23) and (25) can be used to determine the balance between the carrier powers.

6. EXPERIMENTS

Following the results of previous and on-going ESA studies, a signal structure, composed of three carriers in L-band, has been developed. It is expected that this type of signal will allow fast and reliable carrier phase ambiguity resolution as described above, leading to high precision local differential applications (e.g. CAT-III landing and urban navigation with high DOP factors). In the frame of the Experiment Definition Study and as part of the initial trials of Phase I, the technique is to be validated with a signal simulator. If successful, further trials will be conducted as part of the ground experimentation and using ground pseudolites.

However, there will still be various parameters that will affect the performance of the technique in practice (such as tropospheric and ionospheric effects), which will not be fully validated by the ground trials. Therefore, the realisation of an in-orbit flight experiment is fundamental before the three-carrier signal structure can become a realistic candidate for GNSS-2.

The planned in-orbit validation will, for simplicity, make use of a transparent transponder on a geostationary satellite. This does not allow the control of the relative delays between the code and carrier waveforms transmitted by the satellite to meet the requirements for absolute positioning. The demonstration will therefore be made using two-receiver differences. Biases introduced by the lack of coherence will cancel out in the process of taking the differences. It is stressed that the use of a transparent transponder is only intended for experimental purposes and that an operational system would be based on generative payloads.

Scenario of the experiments

Fig. 3 shows the main components of the measurement setup, a (geostationary) satellite transmitter, ground transmitters (pseudolites) and two ground receivers. The latter are referred to in the following as the reference station and the user. Both are used in a static mode, although some experimentation with a mobile user might be envisaged.

The required phase resolution is such that single differencing would imply too stringent requirements on user - reference clock-difference stability. For this reason, double differencing is required, but the experimental scenario contains only one satellite. That is why a supplementary ground transmitter (pseudolite) must be introduced to eliminate user - reference clock differences. In addition, other pseudolites will be deployed in order to provide the necessary minimum number of measurements for supporting the user position determination.

Frequencies

ESA has sent to ITU a request for advance publication of three frequencies to be used for a European Navigation Satellite System (ENSS). These frequencies use parts of the existing allocations in L-band for satellite-to- ground navigation signals. The same frequencies are to be used for the satellite transmissions in the three-carrier experiment.

The experimental frequencies are:

E1 = 1589.742 MHz, (λ_{E1} = 0.189 m)

E2 = 1561.098 MHz, (λ_{E2} = 0.192 m)

E3 = 1215.324 MHz, (λ_{E3} = 0.247 m)

The corresponding difference frequencies and wavelengths are:

E12 = 28.644 MHz, (λ_{E12} = 10.47 m)

and E13 = 374.42 MHz, (λ_{E13} = 0.801 m)

Fig. 4 shows the location of these frequencies with respect to existing satellite navigation systems. The corresponding code chip rates are:

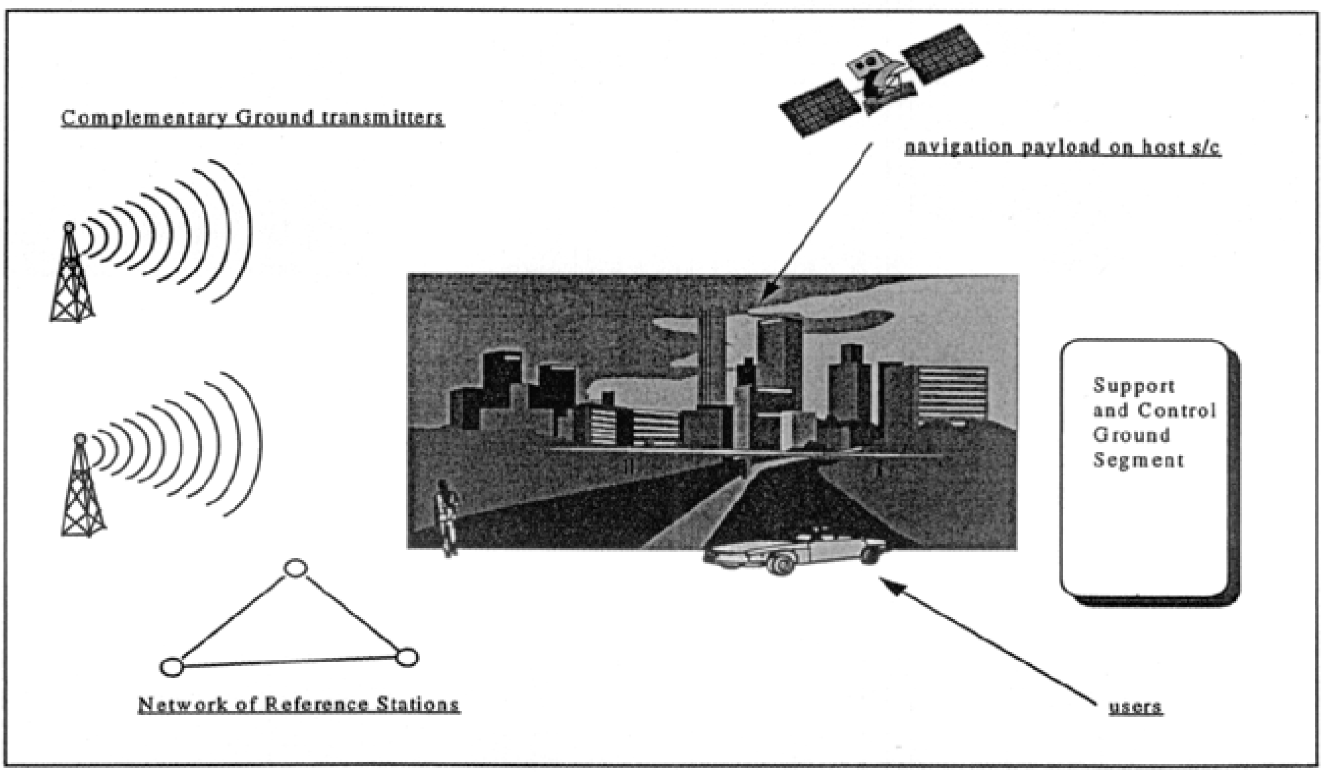

Fig. 3: The Experiment scenario

2.046 Mb/s (λ_{code} = 146.5 m) on E1 and E2,
0.25575 Mb/s (λ_{code} = 1172.2 m) on E3.

Assumptions

Because the double-difference measurements are to be made in a local-area environment with a sufficiently short baseline, the atmospheric and clock biases are assumed to be negligible. It is also assumed that multipath from the antenna surroundings can be reduced to a sufficiently low level.

It is also assumed that the necessary degree of synchronisation between satellite and pseudolite, and between user and reference can be achieved. Requirements in this respect are not very demanding as the use of double differencing cancels clock differences provided that these do not drift significantly during a measurement epoch. It is consequently assumed that thermal noise is the limiting parameter of measurement resolution. However, it should be observed that the dual-frequency method increases the noise of the "ionosphere-free" (i.e. delay compensated) range measurements. This increase is minimised if carriers with the largest frequency spacing are used for all compensations, i.e. in this case E1 and E3. It should also be observed, however, that the low chip rate at E3 implies a low code resolution which also affects the accuracy of the ionospheric correction. For this reason, it may be appropriate to deliver a higher EIRP at that frequency than at the others, to increase the C/N_0 at E3.

Reliability

The overall purpose of the experiment is to find out under what practical circumstances the described sequence of ambiguity resolutions can be carried out with the required probability of success, assuming that the basic principles have been demonstrated in the laboratory.
Two of the most important parameters to be looked at are the distance between reference and user sites, and the signal-to-noise ratio. It has been (tentatively) decided that the required reliability can be obtained by setting $\nu = 4$, meaning that the 4σ error at one stage must be less than half the wavelength at the next one. This will be reviewed in the light of the results of laboratory experiments.

CONCLUSIONS

ESA is engaged in a wide range of studies and experiments in support of the overall goal of establishing a truly international, civilian GNSS. One of the main technical issues is to design the system for high-accuracy use by making carrier phase ambiguity resolution possible in real time. One possible solution is to use three suitably spaced carrier frequencies. The theoretical investigations indicate that this is a feasible method, leading to robust and fairly simple implementation. However, the studies also indicate that there may be limitations with regard to baseline lengths and bias magnitudes. Further investigations are necessary to quantify these. Experiments, both in the laboratory and by means of satellites in orbit, are being prepared.

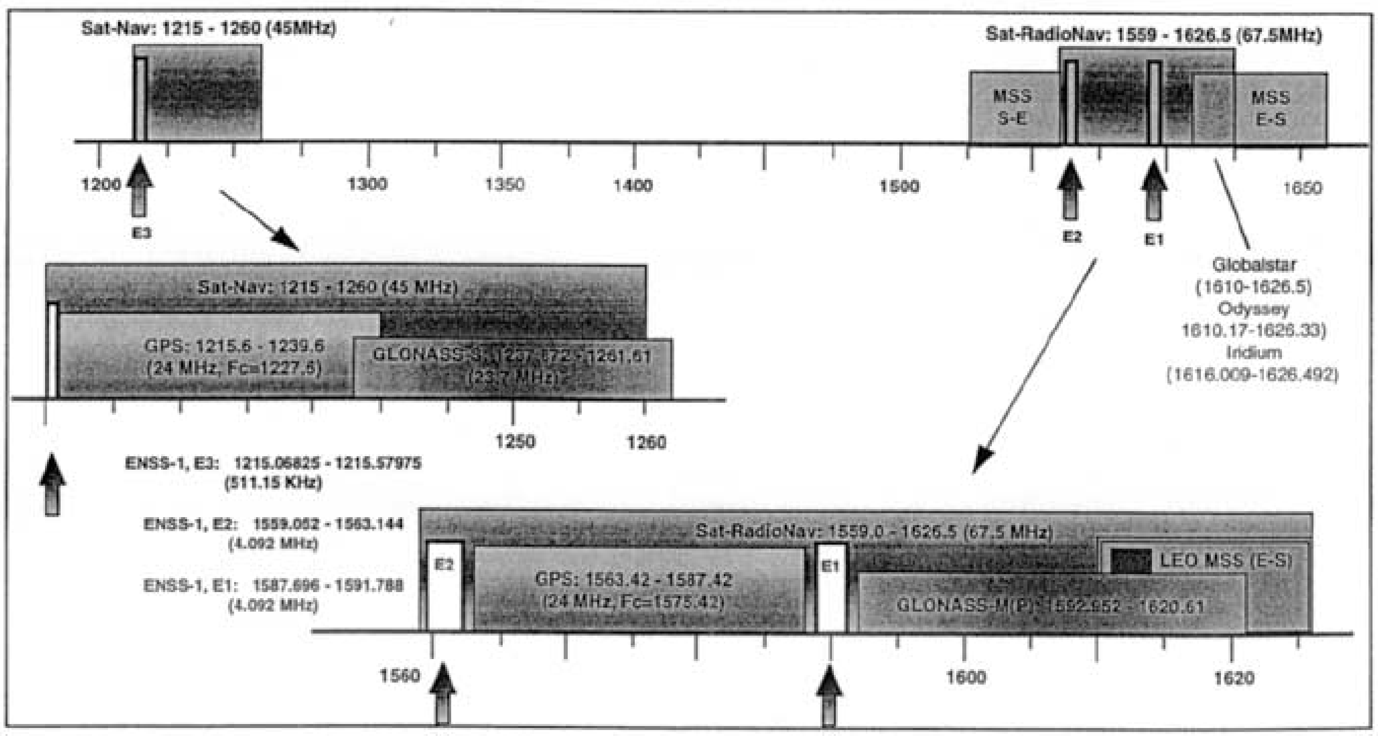

Fig. 4 ENSS Frequencies in Relation to GPS and GLONASS

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