CHAPTER 6

FREQUENCY STABILIZATION

6.1 Introduction

Local oscillators for millimetre-wave radio astronomy receivers designed for molecular spectroscopy must provide a spectrally pure and stable frequency signal. Frequency accuracy and stability of the order of a few parts per billion ($\sim 10^{-8}$) is required for identifying the molecular lines and determining their line profiles. In addition, the frequency of the local oscillator must be switchable at a rapid rate. Frequency switching of the first local oscillator is widely employed in radio astronomy receivers for minimizing the effect of receiver gain fluctuations (Meeks, 1976). It consists of switching the local oscillator frequency between an on-line and an off-line frequency and synchronously detecting the receiver output. Any changes in receiver gain which are slower than the switching rate are cancelled by using this technique. Switching rate is chosen depending on individual receiver gain stability characteristic. However, switching rates between 3 to 400 Hz are most commonly employed. These requirements of the local oscillator source cannot be met by a free-running millimetre-wave oscillator. Therefore, some method of frequency stabilization is needed.

Several frequency stabilization techniques have been developed for millimetre-wave oscillators (Baprawski et al, 1976). These include the use of a high-Q transmission or reaction cavity, an injection-locking technique and phase-lock loops. Use of high

Q-cavities for frequency stabilization of millimetre-wave oscillators is rather limited due to extremely tight dimensional constraints at these wavelengths. Another disadvantage of this scheme is that the stabilized oscillator frequency cannot be tuned.

The injection-locking technique involves injecting a small amount of a stable reference signal into the oscillator output. The oscillator signal is phase-locked to the reference when the frequency of the reference is close to the oscillator output frequency (Adler, 1946). The locking bandwidth is directly proportional to the power of the injected signal and inversely to the oscillator loaded quality factor (Q_L). These characteristics give rise to two main disadvantages of this technique for frequency stabilization of millimetre-wave oscillators. Firstly, a relatively stable auxiliary source of moderate power output is required at the oscillator output frequency, which is difficult and expensive to obtain at millimetre wavelengths. Secondly, the requirement of lower oscillator Q_L for achieving reasonable locking bandwidth leads to the worsening of oscillator FM noise performance.

The phase-lock loop method of frequency stabilization does not suffer from the disadvantages associated with the other techniques. In fact, it combines the advantages of the other techniques because high Q oscillators with good FM noise performance can be phase-locked and yet provide large locking bandwidths and frequency agility.

A versatile phase-lock scheme for the frequency stabilization of millimetre-wave oscillators is described in this chapter. The scheme has been employed successfully in phase-locking several millimetre-wave Gunn oscillators as well as klystrons to a highly stable reference signal derived from a VHF frequency synthesizer. The effect of the phase-lock loop on the oscillator noise characteristics is also discussed.

6.2 Principles of operation of phase-lock loop (PLL)

A phase-lock loop (PLL) basically consists of three components, a phase detector, a voltage controlled oscillator (VCO) and a loop filter. The block diagram of an elementary PLL is shown in figure 6.1. The VCO signal is compared with a reference signal in the phase detector to produce an error voltage which is a function of instantaneous phase difference between the reference and the VCO signal. This error voltage is fed back to the control input of the VCO through a loop filter to close the loop. The VCO frequency is thus continuously corrected till 'phase-locking' takes place. When the loop is locked, the frequency of the VCO is exactly equal to that of the reference. Therefore, a phase-lock loop provides a perfect frequency control wherein the stability of the reference signal is transferred to the VCO.

The dynamic performance of the phase-lock loop is expressed in terms of a natural frequency ω_n and a damping factor ξ which are related to the characteristics of various loop components by the

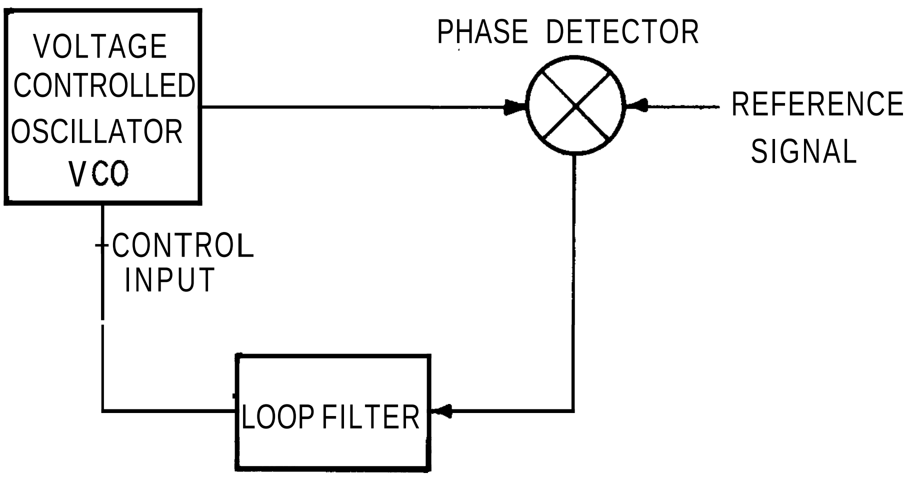


FIG. 6.1 AN ELEMENTARY PHASE-LOCK LOOP.

following relations (Gardner, 1979):

$$\omega_n = \left[\frac{K_o K_d}{\tau_1}\right]^{\frac{1}{2}} \qquad (6.1)$$

$$\xi = \frac{\tau_2}{2}\left[\frac{K_o K_d}{\tau_1}\right]^{\frac{1}{2}} \qquad (6.2)$$

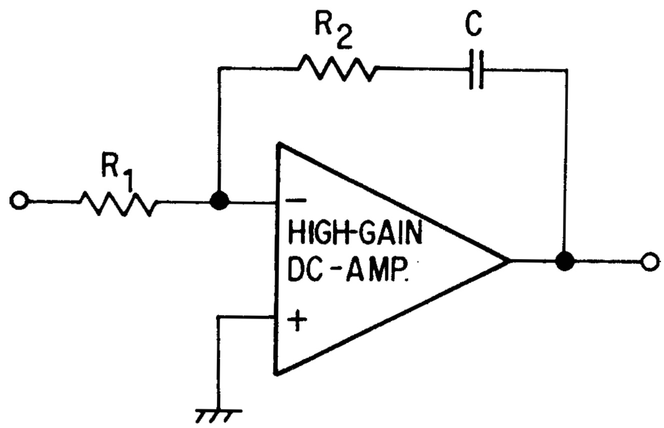
$$\tau_1 = R_1 C \qquad (6.3)$$

$$\tau_2 = R_2 C \qquad (6.4)$$

where K_o is the VCO gain factor expressed in rad/sec-V, K_d is the phase detector gain factor expressed in volts per radian and τ_1 and τ_2 are the time constants of the loop filter shown in figure 6.2. A phase-lock loop employing a loop filter of this type is called a second-order loop by analogy with servo control systems. This is by far the most often used type of PLL due to its excellent performance. Another useful concept for PLL, again borrowed from servo terminology, is the closed-loop transfer function H (s) generally expressed in Laplace Transform domain. The closed loop transfer function of the second order phase-lock loop employing an active loop filter of the type shown in figure 6.2 is given by (Gardner, 1979):

$$H(s) = \frac{2\xi\,\omega_n s + \omega_n^2}{s^2 + 2\xi\,\omega_n s + \omega_n^2} \qquad (6.5)$$

where ω_n and ξ are as defined in equations 6.1 and 6.2 respectively. Equation 6.5 shows that the transfer function of a second order phase-lock loop is low-pass in nature. Therefore, the loop performs



$$\tau_1 = R_1 C$$
$$\tau_2 = R_2 C$$

FIG. 6.2 LOOP FILTER.

a low-pass filtering operation on the reference frequency (phase) input.

ω_n and ξ are optimized for best loop performance. A large value of ω_n gives large loop - bandwidth, but makes the loop unstable. For millimetre-wave klystron stabilization, loop bandwidths around 1 MHz have given good performance (Weinreb, 1970). A low value of ξ (< 1) makes the loop respond quickly while high values of ξ make the loop response sluggish. A ξ value of 0.7 corresponds to a critically coupled loop. However, ξ values around 2.0 have been found to give good loop response consistent with loop stability (Weinreb, 1970).

The practical form of phase-lock loop circuit for frequency stabilization of millimetre-wave oscillators generally contains many more components than those shown in the elementary PLL of figure 6.1. These are necessary for convenience of implementation and do not alter the basic second-order PLL nature of the circuit. A phase-lock loop system developed for frequency stabilization of millimetre-wave oscillators is described in the next section. The system is similar to the one developed by P.S. Henry (1976) for the frequency stabilization of millimetre-wave reflex klystrons.

6.3 Phase-lock system for frequency stabilization of millimetre-wave oscillators

A block diagram of the phase-lock system is shown in figure 6.3.. A sample of the millimetre-wave oscillator signal is mixed

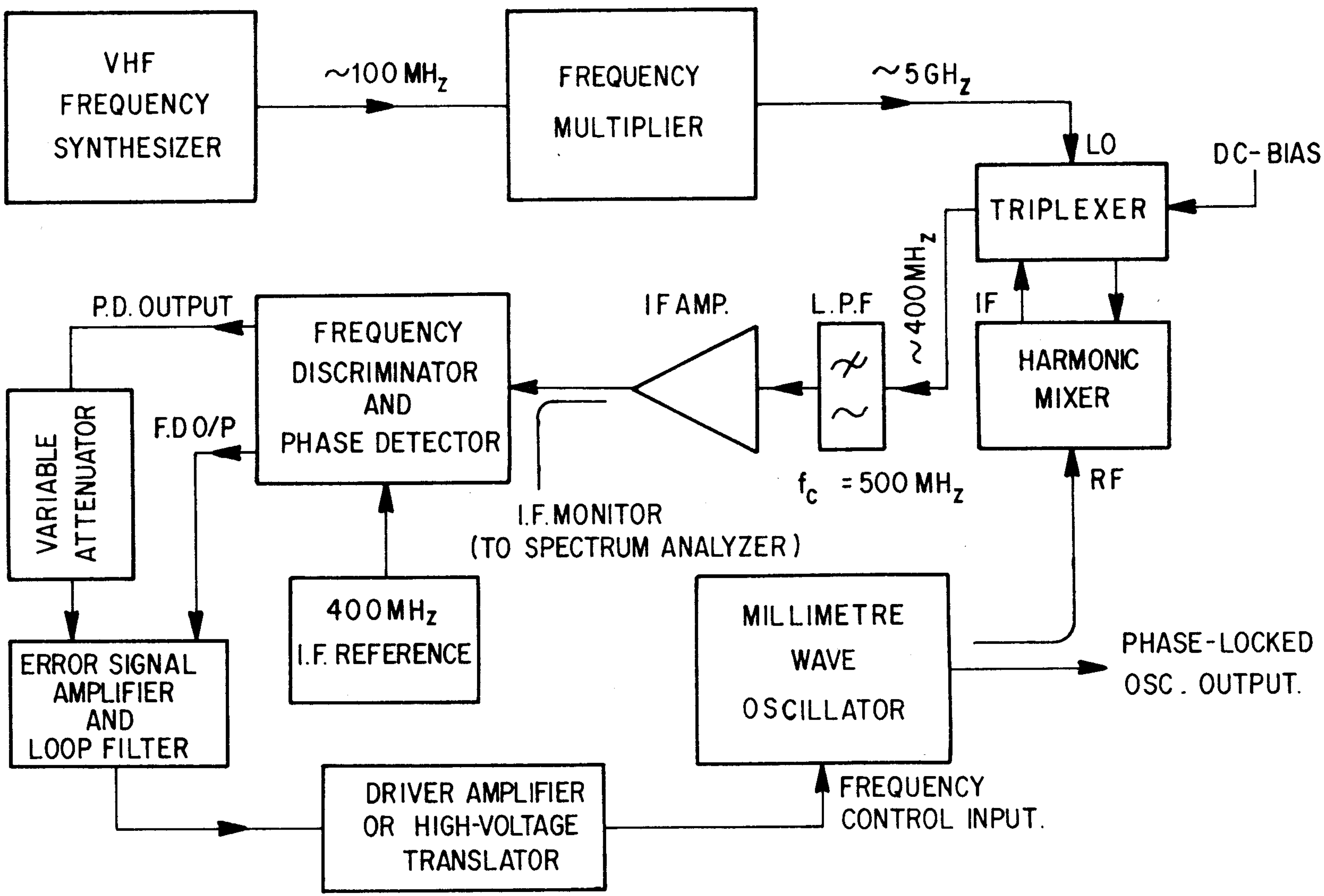


FIG. 6.3 BLOCK DIAGRAM OF THE PHASE-LOCK SYSTEM FOR FREQUENCY STABILIZATION OF MILLIMETRE-WAVE OSCILLATORS.

with the harmonic of a highly stable LO signal around 5 GHz in a harmonic mixer to produce an intermediate frequency (IF) around 400 MHz. The LO signal itself is obtained by multiplying the output of a highly stable VHF synthesizer (Fluke 6160B) in a frequency multiplier (Miteq PLA 4853) which is also based on a phase-lock loop circuit. The IF signal is very weak due to the large conversion loss (∿ 50dB for 20th harmonic of LO) of the harmonic mixer. The low-level IF signal is passed through a low pass filter (L.P.F) with cutoff frequency of 500 MHz to block the leaking LO signal and then amplified in a low-noise IF amplifier. The amplified IF signal is then compared with a stable 400 MHz reference source in a frequency discriminator and phase-detector to produce an error signal which is fed back to the frequency control terminal of the millimetre-wave oscillator to close the loop. In the case of the Gunn oscillators, the error signal is applied through a driver amplifier which also provides the d.c. bias while for klystrons the error signal is applied to the reflector through a high-voltage translator circuit.

The loop locks when the following condition is satisfied:

$$f_o = Nf_{LO} \pm f_{IF} \qquad (6.6)$$

where f_o is the frequency of the millimetre-wave oscillator, f_{LO} is the frequency of the LO signal, f_{IF} is the frequency of the IF signal and N is the harmonic number of the LO signal which mixes with the millimetre-wave signal to produce the IF. A +ve sign is used in equation 6.6 when mm-wave oscillator frequency is higher than the harmonic of the LO and is called upper-sideband (USB) locking while -ve sign is used when mm-wave oscillator frequency is lower than the harmonic of LO and is called the lower-sideband (LSB) lock.

The frequency discriminator is used as an aid for acquisition of lock and gives a d.c. voltage proportional to the frequency difference of the IF signal from the 400 MHz reference. The d.c. voltage shifts the frequency of the millimetre-wave oscillator signal in the appropriate direction for phase-locking. When the frequency difference between the IF and the 400 MHz reference is smaller than the loop bandwidth, the error signal from the phase detector takes over and the loop is locked. When this happens, the frequency discriminator output goes to zero.

A variable attenuator is provided at the output of the phase detector. This is convenient for optimizing the dynamic performance of the loop. The attenuator affects the overall loop gain and hence the locking bandwidth. The attenuator is adjusted to obtain the optimum locked oscillator spectrum while monitoring the IF signal on a spectrum analyzer.

A detailed description of some of the important components used in the PLL scheme is given below.

6.3.1 Harmonic mixer and the triplexer

The harmonic mixer is a relatively simple device. It consists of a Schottky barrier diode mounted in a waveguide. Most mixers available at millimetre wavelengths use a single ended design with only one waveguide input port. The millimetre-wave oscillator signal is directly fed through the waveguide input port. Since the harmonic mixer has only one other (co-axial) port, a triplexer has been built to feed the LO power and dc-bias and to extract the IF output through this port. A circuit diagram of the triplexer is shown in figure 6.4. The triplexer is built on low-loss copper clad substrate utilizing chip capacitors and air core inductors. The insertion loss of the triplexer from the LO input port to the harmonic mixer has been measured to be less than 1dB at 5 GHz while the isolation between LO and IF port is found to be at least 20dB at 400 MHz. DC-bias port isolation with respect to LO and IF ports is greater than 30dB.

6.3.2 Frequency discriminator and phase detector

A block diagram of the frequency discriminator/phase detector circuit is shown in figure 6.5. It is based on the DC Quadricorrelator design used by D. Richman for colour carrier phase synchronization in N.T.S.C. television (Richman, 1954). The IF input is split into two in-phase signals by a two-way power divider and fed to two balanced mixers which also serve as phase-detectors. The reference signal is fed to these mixers after being split into two quadrature components. The output of the quadrature mixer is differentiated and then multiplied with the in-phase mixer output to give the frequency discriminator (FD) output. It can be easily shown (Gardner, 1979) that the output of the multiplier contains a d.c.

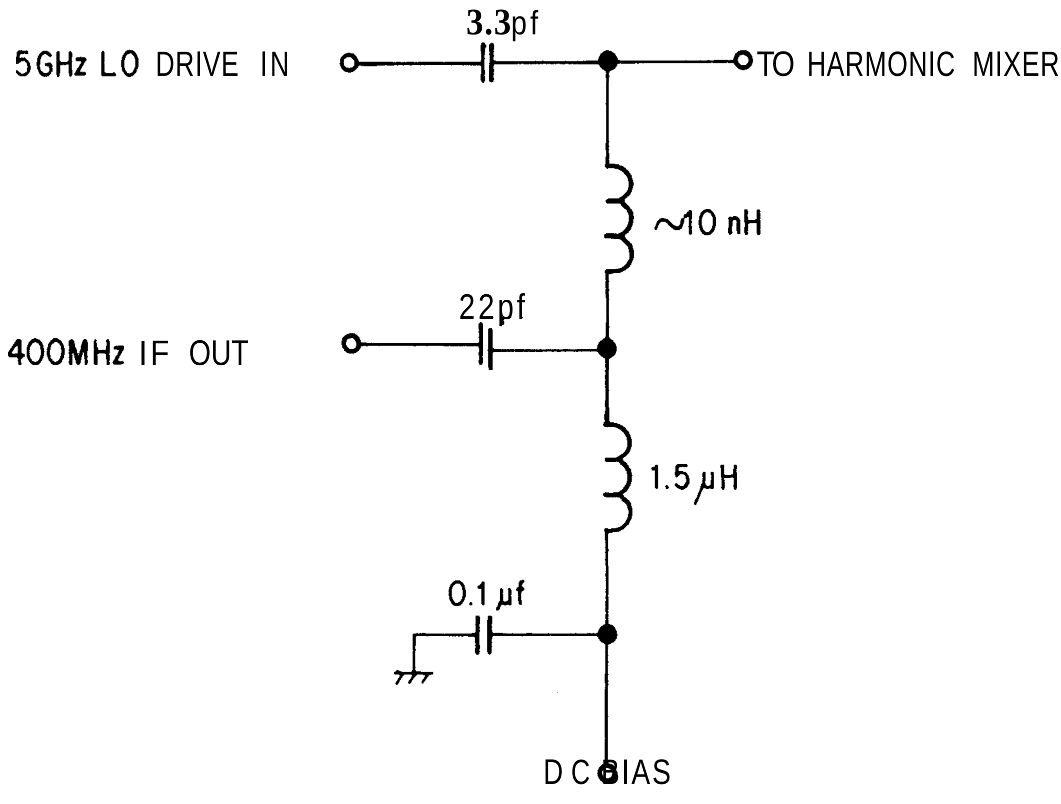


FIG. 6.4 CIRCUIT DIAGRAM OF THE TRIPLEXER FOR THE HARMONIC MIXER.

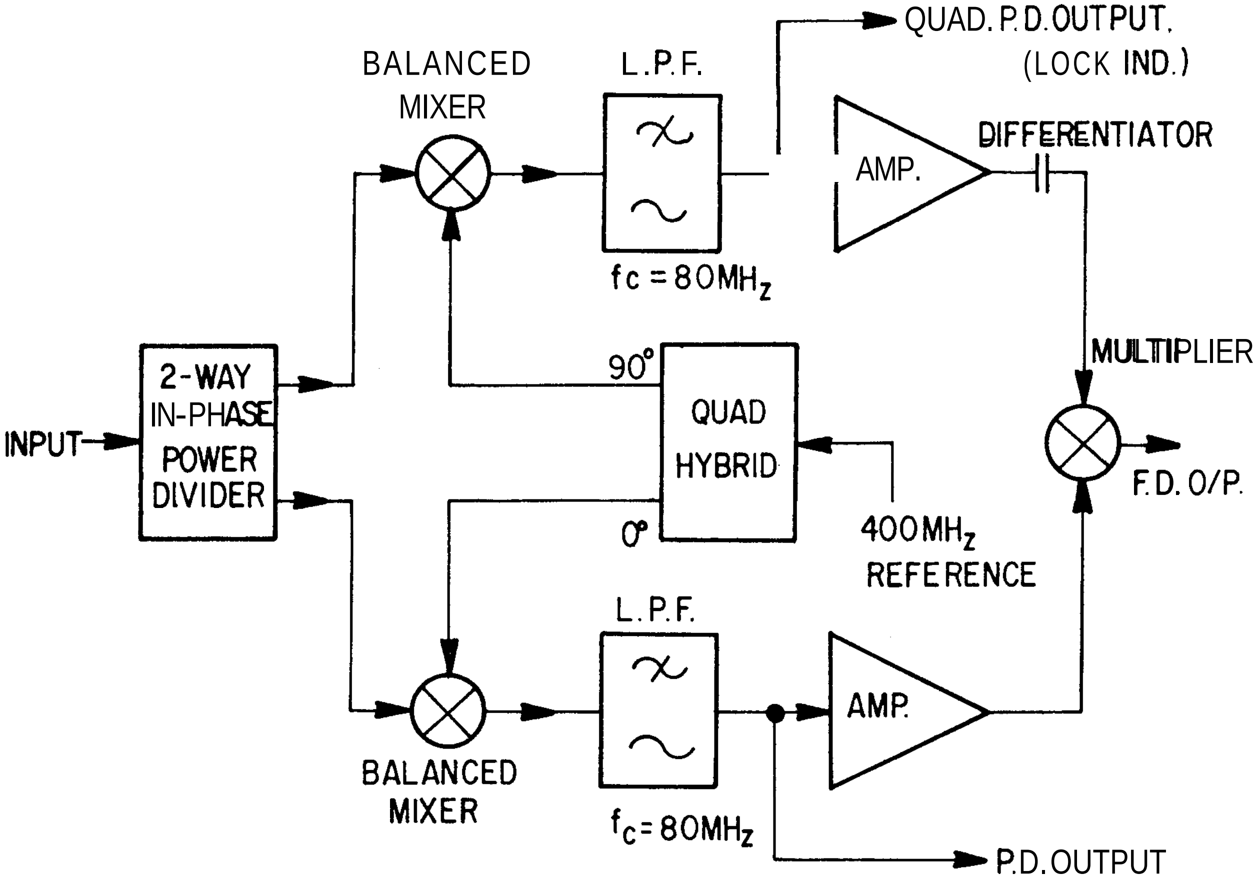


FIG. 6.5 BLOCK DIAGRAM OF THE FREQUENCY DISCRIMINATOR/PHASE DETECTOR.

component proportional to the frequency difference between the IF and the reference signals. The in-phase mixer also serves as the loop phase-detector (PD) while the quadrature mixer output is utilized for lock indication. When the loop is locked, the d.c. output of the in-phase mixer is nearly zero while the quadrature mixer output is maximum. Frequency-voltage characteristics of the discriminator are shown in figure 6.6 when the IF input is swept from 275 to 525 MHz. In the phase-lock system the discriminator is used between 320 and 480 MHz, over which range the discriminator output is fairly linear. The IF input level is kept at +3dBm and the reference level at +10dBm while obtaining the characteristics of figure 6.6.

6.3.3 Error signal amplifier

An error signal amplifier is used for processing of the error signal before being applied to the oscillator frequency control input. A circuit diagram of the error signal amplifier is shown in figure 6.7. The frequency discriminator (FD) output is amplified in a d.c. amplifier using a low drift op-amp AD517LH and added on to the phase detector (PD) output signal in a summing amplifier using a high frequency op-amp AD50J. Since the discriminator output voltage polarity depends on whether the IF frequency is below or above the reference, a polarity reversal switch (LSB/USB selector) is also provided in the d.c. amplifier to select the correct polarity for negative feedback. The summed output signal (FD+PD) is processed through two parallel paths. A d.c. path which goes through a driver amplifier wherein the d.c. bias voltage of the Gunn oscillator is also

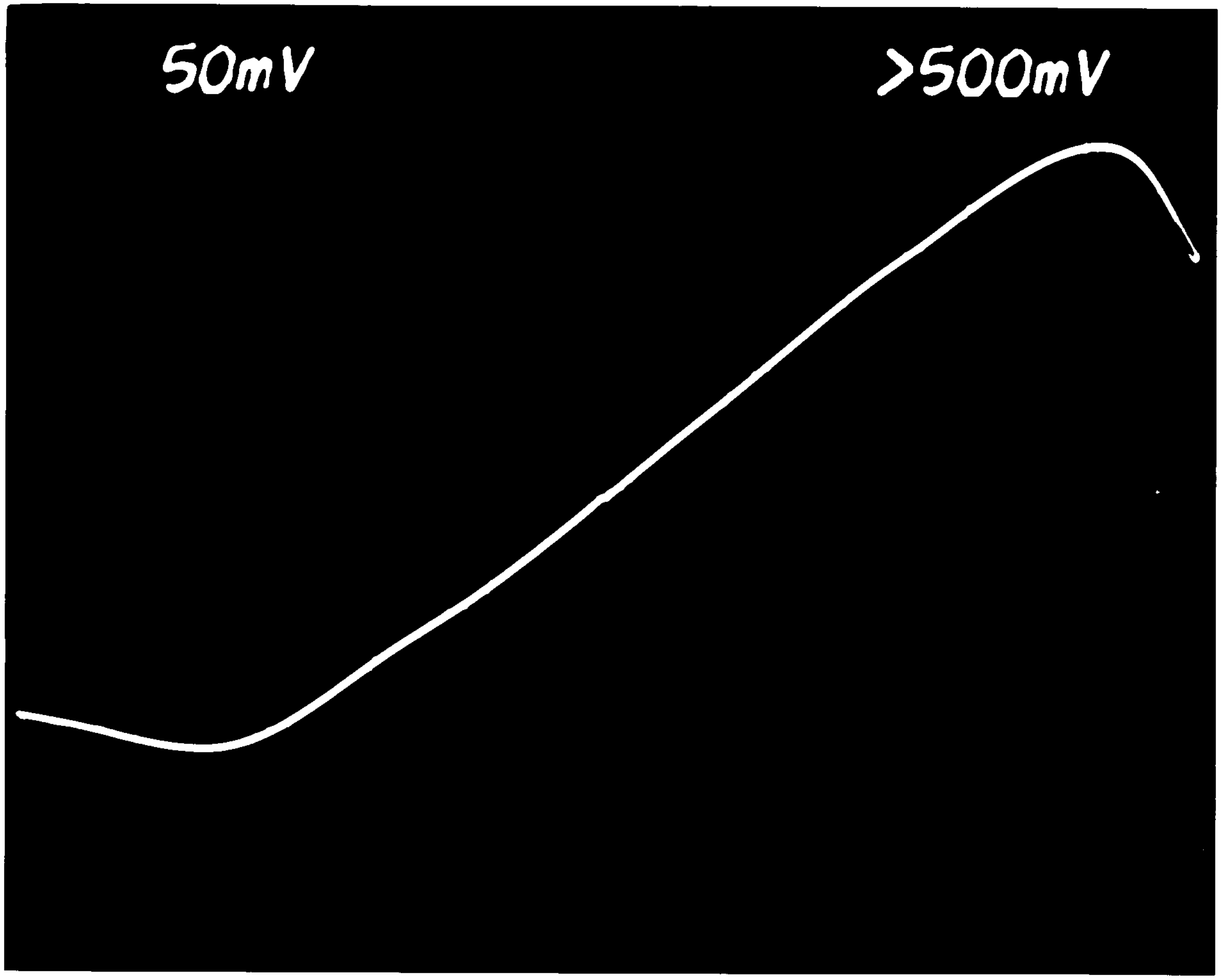


FIG. 6.6 VOLTAGE-FREQUENCY CHARACTERISTICS OF THE FREQUENCY DISCRIMINATOR FOR I.F. INPUT FREQUENCIES BETWEEN 275 to 525 MHz. HOR-25 MHz/div., VER-50mV/div.

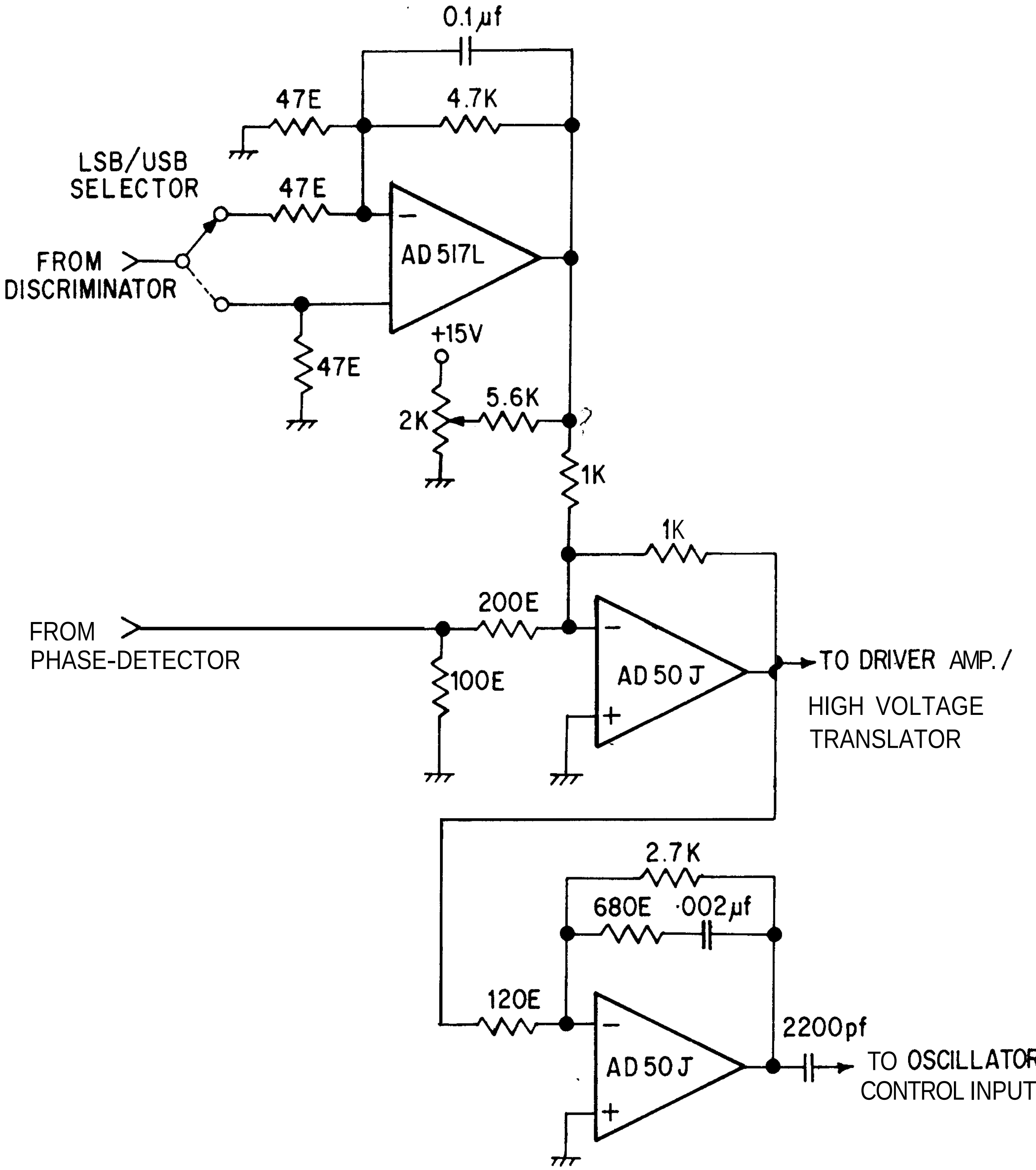


FIG. 6.7 CIRCUIT DIAGRAM OF THE; ERROR SIGNAL AMPLIFIER,

added on to the error signal, and a high frequency a.c. path where the error signal is capacitively coupled directly to the oscillator frequency control input. A loop filter of the type shown in figure 6.2 is also included in this path as shown in the figure. The two paths together provide a maximum loop bandwidth of about 5 MHz. The actual loop bandwidth is optimized using the attenuator at the output of the phase detector and may be anywhere between 1 and 5 MHz. AD50J op-amps are used for the summing amplifier as well as the loop filter due to their extremely fast response time (Slew-rate 500V/μsec).

6.3.4 Driver amplifier [for Gunn oscillators)

The d.c. bias requirement of millimetre-wave Gunn oscillators ranges from 3 to 5 volts with a maximum operating current of around 2 amperes. This relatively large d.c. drive must be provided by the loop amplifier for bias-tuned oscillators. A driver amplifier used for the purpose is shown in figure 6.8. The d.c. bias is added to the error signal using a high frequency op-amp (TP1321) summing circuit. The output of the op-amp is used to drive a high frequency power transistor (Motorola type 2N5646) which provides the required operating current for the Gunn oscillators. An n-p-n transistor 2N2222 is used to isolate the op-amp output from the power transistor and to provide the necessary base-drive for the power transistor.

6.3.5 High voltage translator (for klystrons)

The frequency of millimetre-wave reflex klystrons is electronically tunable by charging the reflector voltage. The normal operating voltage for the reflector is of the order of -2 to -3KV with respect

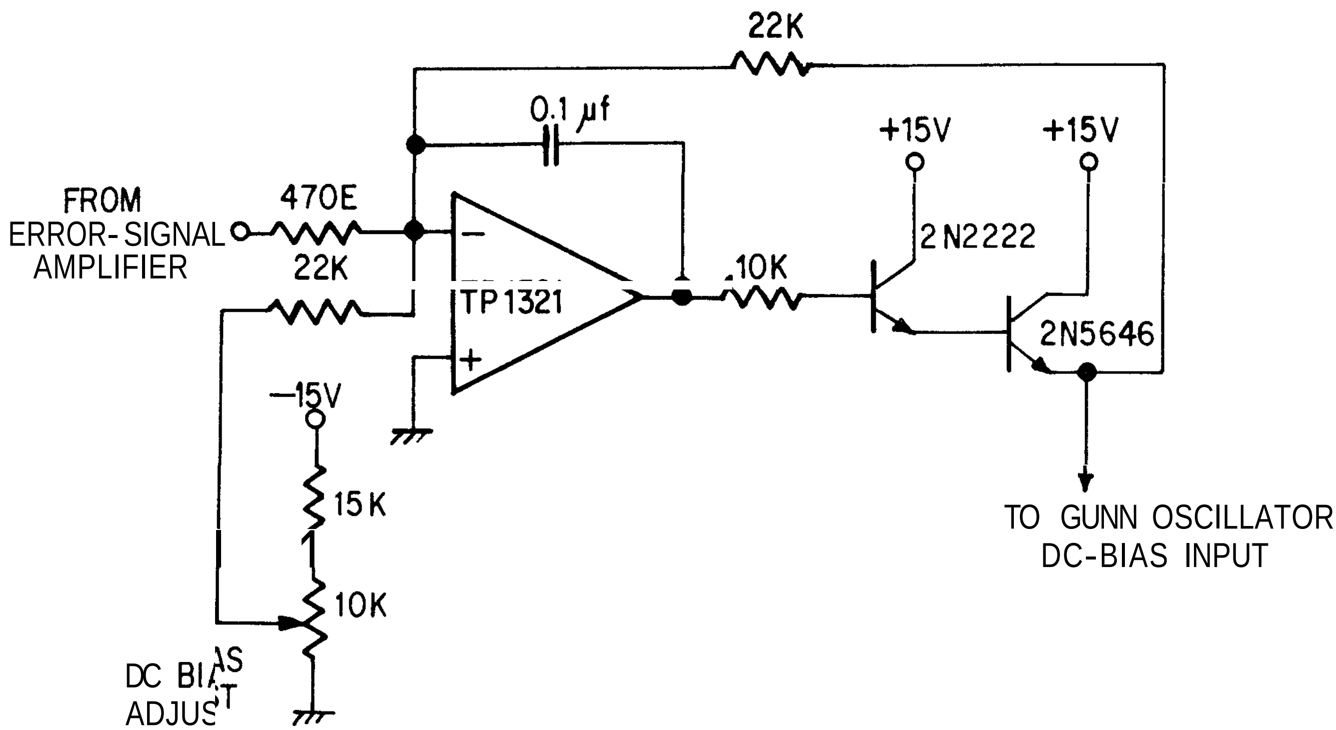


FIG. 6.8 CIRCUIT **DIAGRAM OF THE DRIVER** AMPLIFIER.

to ground. For phase-locking these klystrons using the phase-lock scheme described here, the error signal must be translated from near the ground potential to the very high voltage present at the reflector terminal. In addition, the error signal amplifier must be electrically isolated from the high voltage supply. A scheme used for high voltage translation of the error signal is shown in figure 6.9. Two parallel paths, a d.c. path and a high frequency a.c. path are used here also to couple the error signal to the klystron reflector input. The d.c. path coupling is achieved using a high isolation voltage opto-isolator (Optron OPI 120 with 15KV isolation). The error signal is fed to the input light emitting diode of the opto-coupler while its output photo-transistor is operated with d.c. regulated supplies floating on the reflector voltage. The output of the photo-transistor is further amplified in a high voltage operational amplifier (AD171J) to achieve error-signal voltage levels suitable for the klystron reflector. The high voltage op-amp also operates on floating d.c. supplies. The floating d.c. regulated power is obtained by using a special power-line transformer with 5KV isolation between primary and secondary windings as shown in figure 6.10. The secondary of the transformer floats on the klystron reflector voltage. A ± 10V regulated supply for opto-isolator photo-transistor is obtained using standard low voltage regulators while ± 100V required for the high voltage op-amp is obtained using zener diodes.

The a.c. path coupling is provided directly using a high break-down'voltage capacitor (2200 pF, 6KV). This capacitor along with the

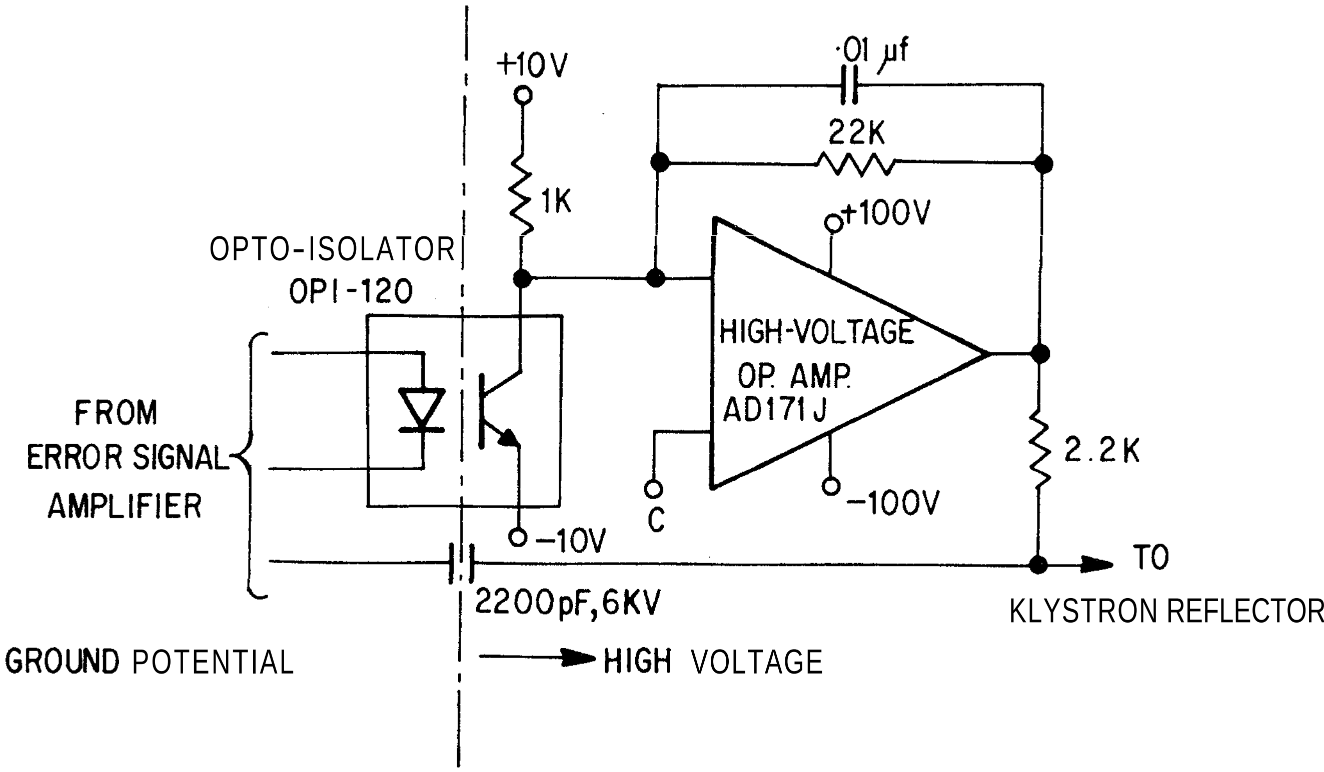


FIG. 6.9 CIRCUIT DIAGRAM OF THE HIGH VOLTAGE TRANSLATOR FOB MILLIMETRE WAVE REFLEX KLYSTRONS.

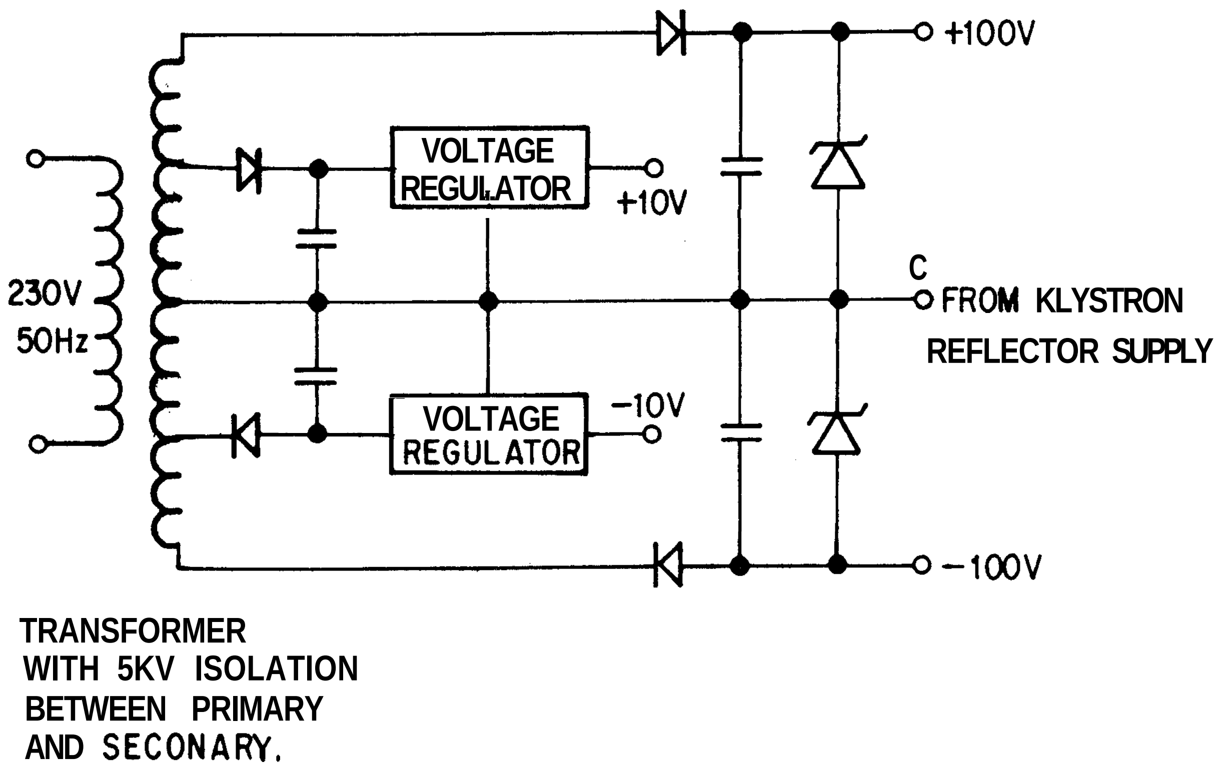


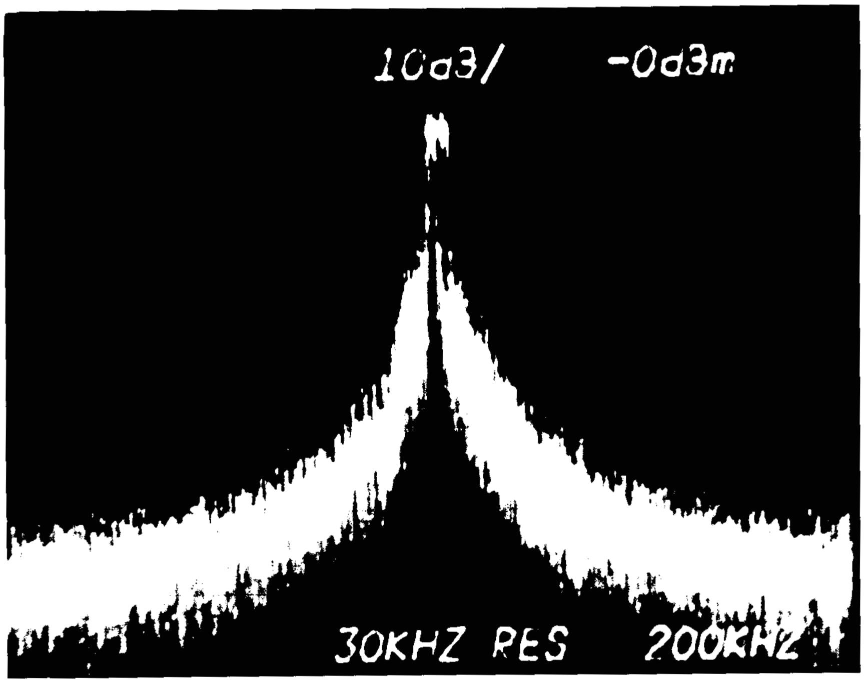
FIG. 6.10 CIRCUIT DIAGRAM OF THE FLOATING d.c. REGULATED POWER SUPPLIES.

resistor at the output of the high voltage op-amp determine the overall frequency response of the circuit.

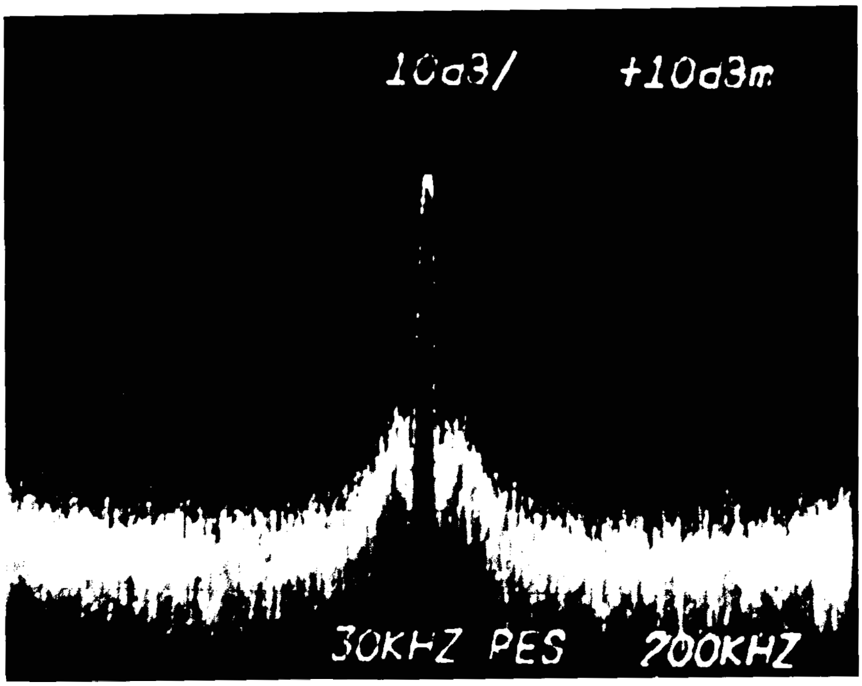
6.4 Performance of the phase-lock system

The phase-lock system described in the foregoing sections has been used to phase-lock a variety of millimetre-wave oscillators in different frequency bands. In all cases, the loop successfully phase-locked the millimetre-wave oscillator signal to the frequency synthesizer output. Slight trimming of the variable nttenuator after the phase detector (refer figure 6.3) is the only adjustment required when millimetre-wave oscillators are changed. This is to compensate for the different voltage-tuning sensitivities of the oscillators. Spectrum analyzer displays of a free-running and phase-locked 87 GHz millimetre-wave Gunn oscillator obtained by monitoring the IF signal are shown in figure 6.11. The stability improvement due to phase-locking is obvious from the display. The effect of an improper setting of the variable attenuator is illustrated in figure 6.12 which shows the locked-oscillator output spectra far too-high, optimum and too-low values of the loop gain. Too-high a value of loop gain makes the loop unstable while too-low a value makes acquisition of lock difficult. The optimum setting roughly corresponds to the almost flat noise side-bands as seen in the oscillator output spectrum of figure 6.12(b).

Several millimetre-wave Gunn oscillators for the 33-50 GHz and 75-110 GHz frequency bands described in chapters 3 and 4 have been successfully phase-locked using this phase-lock scheme. In addition,

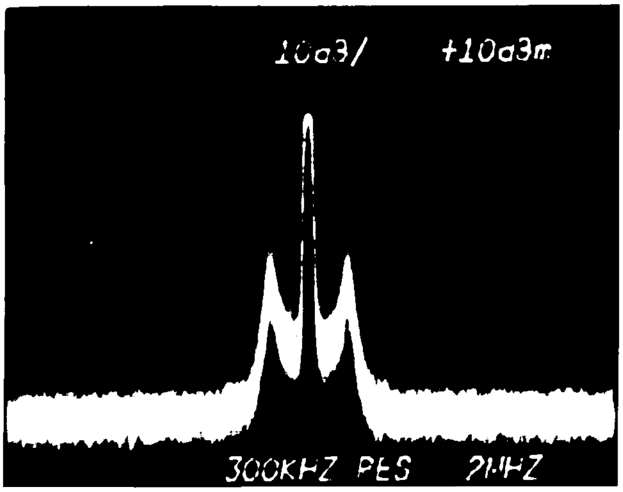


(a)

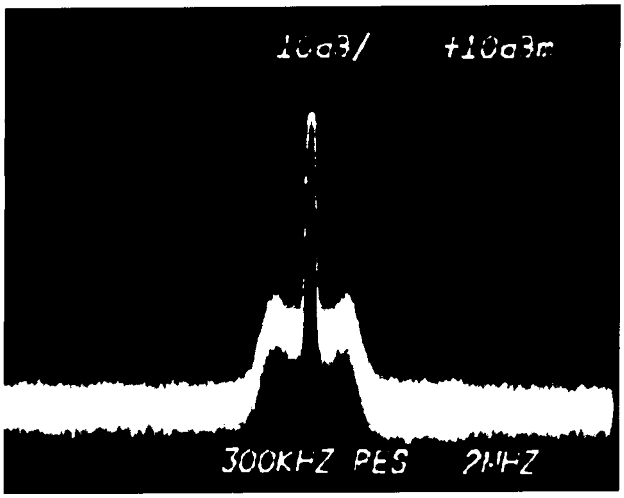


(b)

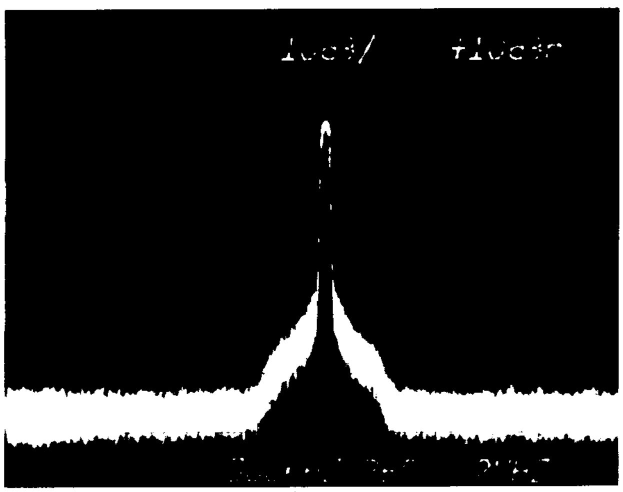
FIG. 6.11 OUTPUT SPECTRA OF AN 87 GHz GUNN OSCILLATOR. (a) FREE-RUNNING, (b) PHASE-LOCKED. HOR-200 KHz/div., VER-10dB/div.



(a)



(b)



(c)

FIG, 6.12 OUTPUT SPECTRA OF THE 87 **GHz** PHASE–LOCKED **GUNN OSCILLATOR** ILLUSTRATING THE EFFECT OF LOOP GAIN (a) LOOP GAIN TOO HIGH the same system with minor variations has also been used for phase-locking a 21 GHz Gunn oscillator meant for local oscillator application in a low-noise radio astronomy receiver designed for water-vapour line (22.235 GHz) observations (Arora and Sarma, 1982). A 110 GHz reflex klystron has also been phase-locked using the same system along with the high voltage translator described in section 6.3.5. The frequency of the locked oscillators can be switched by about ± 80 MHz by switching the VHF synthesizer frequency. The lock acquisition time has been measured to be about lmsec. This is of the same order as the switching time of the VHF synthesizer (∿ 0.8msec) itself.

6.5 Effect of phase-lock loop on oscillator noise

The phase-lock loop does not affect the AM noise spectrum of the millimetre-wave oscillator in any significant manner, but modifies its FM noise spectrum as per the following relation (Baprawski et al, 1976):

$$S(\omega)_o = S(\omega)_{osc} \left|1-H(s)\right|^2 + \left[S(\omega)_R + N^2 S(\omega)_{LO}\right] \left|H(s)\right|^2 \qquad (6.7)$$

where $S(\omega)_o$ is the FM noise spectrum of the locked millimetre-wave oscillator, $S(\omega)_{osc}$ is the FM noise spectrum of the free-running mm-wave oscillator, $S(\omega)_R$ is the FM noise spectrum of the 400 MHz reference oscillator, $S(\omega)_{LO}$ is the noise spectrum of the multiplied synthesizer output, $H(s)$ is the closed loop transfer function of the PLL, and N is the harmonic number of the LO signal which mixes with the millimetre-wave signal to produce the IF. The additive noise due to PLL circuit components is not included in equation 6.7.

Since the closed loop transfer function of a PLL is low-pass in nature, the FM noise of the reference oscillator and the synthesizer signal is low-pass filtered and added to the millimetre-wave oscillator output. The FM noise of the free-running millimetre-wave oscillator, on the other hand, undergoes a high pass filtering operation by the action of the phase-lock loop.

The FM noise contribution of the LO signal (derived from the VHF frequency synthesizer) to the millimetre-wave oscillator output is enhanced by a factor of N^2 due to the N times multiplication of this signal in the harmonic mixer. The enhancement of FM noise in multiplication process comes about as a result of an increase of the frequency deviation of the noise-modulated signal by the multiplication factor. Since the power in FM noise sidebands is proportional to the square of frequency deviation, the factor of N^2 in equation 6.7 is obtained when the L.O. signal undergoes a multiplication by a factor of N in the harmonic mixer. This puts severe constraints on the FM noise characteristics of the VHF synthesizer for application in this type of phase-lock system. However, this is not a serious limitation, since high quality frequency synthesizers with extremely low FM noise are readily available at VHF frequencies (e.g. Fluke 6160B).

6.6 Discussion

The phase-lock system described in this chapter for the frequency stabilization of millimetre-wave oscillators is highly versatile. It can be easily adapted for use with oscillators of different frequency bands by simply changing the harmonic mixer and the directional coupler, the only two waveguide components used in the system. 400 MHz IF comparison frequency is chosen due to the specific system requirements of the millimetre-wave radioastronomy receiver. However, IF comparison frequencies anywhere between 30 MHz - 1000 MHz can be employed. The choice of IF comparison frequency is also determined by the availability of high quality RF processing components. In principle, the VHF synthesizer output itself can be used as the IF reference also, but a separate IF reference source gives greater flexbility of operation, particularly in frequency switching applications. The noise requirement on this source is very moderate since the noise contribution of this source to the overall noise can almost be neglected as compared to the N^2 times enhanced noise of the LO signal (refer eq. 6.7).

The same phase-lock system has been utilized for frequency stabilization of millimetre-wave klystrons as well as Gunn oscillators with only minor modifications in the error signal processing circuit. The use of a frequency discriminator for lock acquisition in place of a search voltage sweep commonly employed in conventional PLL circuits gives considerably improved performance particularly in frequency-agile applications.

CHAPTER 7

CONCLUSION

The design and development of low-noise solid-state sources for local oscillator application in millimetre-wave radio astronomy receivers have been discussed in this thesis. These sources have been developed using commercially available packaged GaAs Gunn diodes. Gunn oscillators for the 33-50 GHz frequency band, an atmospheric transmission window suitable for ground based radio astronomical observations, have been realized in several waveguide circuit configurations. Post-coupled Gunn oscillator in the standard rectangular waveguide circuit gives 50-100mW of CW power at any frequency between 33-47 GHz using a Varian Gunn diode type VSQ-9219S4 which is specified for 75mW output at 43 GHz. The diameter of the bias-post has a strong influence on the oscillation frequency and power output of this type of oscillator circuit. A frequency saturation phenomenon in oscillator backshort tuning has been observed which limits the mechanical frequency tuning obtainable to within 5%. A wideband backshort-tunable Gunn oscillator has been realized using a reduced-height waveguide circuit. This circuit gives 40-80mW output powers over the frequency range 38-44 GHz using the same Varian Gunn diode as described earlier. More than 15% mechanical tuning has been obtained by varying the position of the backshort. A post-coupled Gunn oscillator in a novel circuit configuration using circular waveguide has also been developed. More than 80mW CW power over the frequency range 43-45 GHz has been obtained with this circuit using a Varian 45 GHz-100mW Gunn diode (type VSQ-9219S5). The circular waveguide circuit is comparatively simple in construction but requires a circular to rectangular waveguide transition for use in the standard waveguide systems. This, however, is not a serious disadvantage since such transitions are commercially available and quite inexpensive.

Gunn oscillators for the 75-110 GHz frequency band, another atmospheric transmission window region, have been realized in a resonant-cap circuit suitable for harmonic extraction. A new design using a resonant-cap circuit in a circular waveguide has been developed which is relatively simpler to construct at millimetre wavelengths compared to the standard rectangular waveguide circuits. An experimental investigation of the effect of various cap dimensions on the oscillation frequency has shown that the oscillator frequency is a well behaved function of the resonant-cap dimensions. An empirical relation has been obtained for the determination of the oscillation frequency from the resonant-cap dimensions. The performance of the circular waveguide Gunn oscillator has been compared with that of a conventional resonant-cap circuit realized in a standard rectangular waveguide. The output powers and the mechanical tuning range obtained with the new circuit are comparable to that of the conventional design. 10-20mW CW powers have been obtained over the frequency range 75-90 GHz using a Varian Gunn diode type VSB-9222S2 specified for 10mW output at 87 GHz. Mechanical tuning range of about 8 GHz within 3dB variation in output power has also been achieved by varying the position of the resonant-cap Gunn diode assembly across the waveguide. The oscillator power is seen to fall-off rather sharply above 95 GHz, perhaps due to the frequency limitation of the GaAs Gunn diode itself.

The output powers obtained from the Gunn oscillators are more than adequate for local oscillator application in radio astronomy receivers. AM sideband noise at 1.4 GHz (I.F. of the millimetre-wave radio astronomy receivers) away from the carrier which is a critical parameter for local oscillators, has been measured for several Gunn oscillators in the 75-110 GHz band. Noise-to-carrier ratio of about -145dBc in 1 KHz bandwidth obtained for the Gunn oscillators is 6-12dB lower than that obtained with some commercially available millimetre-wave reflex klystrons.

The frequency stabilization of Gunn oscillators; an important local oscillator requirement for radio astronomy receivers designed for molecular spectroscopy, has been achieved using a versatile phase lock loop circuit. The phase-lock circuit translates the extremely high frequency stability of a VHF frequency synthesizer output to the millimetre-wave Gunn oscillator. The circuit has successfully been used to phase-lock several millimetre-wave Gunn oscillators as well as klystrons to the frequency synthesizer output.

Solid-state local oscillators using Gunn diodes have thus been realized for providing a reliable alternative to the highly expensive and short-lived klystrons at least up to 95 GHz. Millimetre-wave frequency multipliers (Archer, 198.1; Erickson, 1982) may be required to provide solid-state local oscillator power beyond 95 GHz. These multipliers employ whisker contacted Schottky diode varactors in a crossed-waveguide mount and typically give conversion efficiencies of 5-20% for output frequencies in the 100-300 GHz range. These multipliers can be driven with lower frequency Gunn oscillators for an all solid-state local oscillator realization (Archer, 1982). On the other hand, InP Gunn diodes have shown good promise for efficient operation and excellent noise characteristics beyond 100 GHz. These may provide the next generation of solid-state millimetre-wave local oscillator sources (Eddison and Davies, 1982). However, a number of problems in growing InP epitaxially remain to be solved before these devices become commercially available. Therefore, GaAs Gunn diodes along with the Schottky diode multipliers may continue to provide solid-state low-noise millimetre-wave power for local oscillator application for some more time.

There is considerable scope for theoretical work on the analysis of millimetre-wave Gunn oscillators based on the large amount of experimental data provided here. A theoretical understanding of the oscillator behaviour should lead to further improvements in the oscillator circuit design. Advances in semiconductor material growth technology as well as device packaging may also be required to complement the improvements in oscillator circuit design for the realization of solid-state millimetre-wave sources of even better performance.

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