CHAPTER - IV
4.0 SOME MICROWAVE SUBSYSTEMS

Compactness, reliability, efficiency and ability to work at varying temperature environment are important parameters of the subsystems for air borne and space based applications. Compact systems are designed by accommodating more and more functional blocks in a single module, as an example, is the active integrated antenna [72], [73]. In this case, active antenna is designed integrating active circuits such as amplifiers, mixer and filters within the same module. Compactness of the circuit not only reduces the size and weight of the system but also improves the system efficiency eliminating interconnection losses and also improves reliability due to the elimination of various interface cables and connections. To operate the complete system in the varying temperature environment, keeping in mind the above important parameters, design aspect of the different temperature compensation mechanism has to be explored. The proper temperature compensation mechanism reduces the complex heat sink requirement to maintain stringent temperature limits of the circuits and systems. The thesis work discussed the temperature behavior and compensation mechanism of four spacecraft subsystems, first is diode based vector modulator for beamforming network [70], [71] second is diode-based linearizer [39], [40], another is, over-drive level controlled microwave solid-state high power amplifier [80] and another is a Ku-band solid-state channel amplifier with automatic level control system [83]. These subsystems contain GaAs FET, HEMTs, PIN diode and Schottky barrier diode. Temperature compensation schemes for these subsystems are proposed and designed. Mathematical and practical design procedures are also developed and presented to determine the circuit component values.
4.1 A Temperature Compensated Diode-Based Vector Modulator for Beam Forming Network

Phased-array antennas are gaining significance for on-board satellite applications because of their ability to form multiple beams and to provide power sharing among beams [62] – [71]. These phased array antennas offer improved operational flexibility by providing independent beam reconfigurability and steerability, resulting in more efficient use of satellite-power resources. Moreover, with the generation of narrow beams, higher effective isotropic radiated power (EIRP) can be achieved, enabling direct communication with small earth stations.

Important consideration in selecting the beamforming structure is the kind of array and the beamforming function implementation. Orthogonal beamforming based on crosobar arrangements is well-known technique particularly efficient when the matrix elements are designed to control the relative phase and amplitude by command [69]. These kinds of matrix elements require circuits providing both amplitude and phase control. These control circuit must be of small size, lightweight, of low power consumption. Also amplitude and phase characteristic of the network must be temperature invariant to operate it wide temperature varying space environment.

In general beamforming network comprising different individual circuits to perform the amplitude and phase-control tasks: variable phase shifters, variable gain/attenuation stages and switches. This approach results in higher insertion losses and cumulative errors in combined results, larger size, and more complicated housing than using a single circuit.

In this section, an single circuit approach vector modulator using PIN diodes with novel optimum load-line biasing scheme for temperature invariant performance is presented [70], [71]. The circuit is realized in a balanced configuration with very good input and output return loss over the operating frequency range of 2.5 to 2.7 GHz. for the entire amplitude (up to -22 dB) and phase variation (360°). Other two vector modulators are also realized at satellites C-band and Ku-band down link frequency.

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4.1.1 Design Approach

A schematic block diagram of the circuit is shown in Figure-4.1.1 [70]. Here input signal is divided into two parts with equal amplitude but with phase difference of 90° with the help of a 3-dB 90° power divider. Then both the signals are selectively attenuated and phase shifted by an amount of 0° or 180° using voltage/current controlled variable attenuator and finally combined by an in-phase power combiner. With this design approach, all possible amplitude and phase combination can be achieved as shown in the phaser diagram as shown in the figure.

![Block schematic of the vector modulator](image)

Fig.-4.1.1: Block schematic of the vector modulator
All the circuits are realized on 25mil alumina ($\varepsilon_r = 9.9$) substrate. The circuit is simulated and analyzed using Series-IV software of HP-EEs of.

### 4.1.2 Details of Circuit Design

Input 3-dB 90° coupler is implemented by Lange coupler to achieve good amplitude and phase performance over broad frequency band. Output 3-dB in-phase power combiner is realized using Wilkinson type power divider.

The most important circuit module is the variable attenuator with all possible attenuation range and with the provision of 180° phase switching. The module is realized using reflection type PIN diode attenuator. Schematic of the attenuator circuit is shown in Figure-4.1.2. 3-dB 90° Lange coupler is used to realize variable attenuator. Pair of PIN diodes are connected to the direct and coupled port of the Lange coupler and isolated port used as output port of the variable attenuator. The basic working principle of the circuit is that $S_{11}$ of the diode terminated port will be the $S_{21}$ of the circuit. The incoming signal at the input port is split equally between the direct and coupled ports, with a 90° phase difference. A further 3-dB split and 90° phase difference is introduced when the waves are reflected back through the coupler. When the two reflected signals are superimposed at the input port they are in antiphase and thus cancel. At the isolated port of the coupler the two reflected signals are in-phase and thus reinforce each other to form the output signal. Thus for Lange coupler of characteristic impedance $R_0 (= 50\Omega)$,
$S_{21}$ of the attenuator circuit, which is the reflection coefficient ($\rho$) of the diode terminal, is given by:

$$S_{21}^{(Att)} = \frac{Z_D - R_D}{Z_D + R_D} \quad (4.1.1)$$

Where, $Z_D$ is diode impedance. Hence a pure variable resistance of diode would lead to a variable $S_{21}$ amplitude, achieving both positive and negative signs. Therefore, $S_{21} = 0$ when the impedance of the diode will be $50$ $\Omega$. Considering diode impedance as pure variable resistor, the amplitude and phase of $S21$ of the variable attenuator circuit is shown in Figure-4.1.3.

Equivalent circuit of a forward biased PIN diode can be represented by a voltage variable resistor $R_d$ with parallel combination of a capacitor $C_d$ as shown in Figure-4.1.4 [27]. Thus considering Lange coupler as a ideal 3-dB 90° coupler, $S_{21}$ of the circuit will be given by (4.1.2):

$$S_{21}^{(Att)} = \frac{\left( R_d \parallel j\omega C_d \right) - R_o}{\left( R_d \parallel j\omega C_d \right) + R_o} \quad (4.1.2)$$

$\fig 4.1.3$: $S_{21}$ variation on diode resistance

$\fig 4.1.4$: RF equivalent circuit of forward biased PIN diode
Thus amplitude and phase of the variable attenuator circuit is given by:

\[ |S_{21}|_{(\text{An})} = \left[ \frac{(R_d - R_o)^2 + (\omega C_d R_d R_o)^2}{(R_d + R_o)^2 + (\omega C_d R_d R_o)^2} \right]^{\frac{1}{2}} \]  (4.1.3)

\[ \angle S_{21} = \tan^{-1}\left[ \frac{2\omega C_d R_d^2 R_o}{R_d^2 - R_o^2 - (\omega C_d R_d R_o)^2} \right] \]  (4.1.4)

The circuit is realized by using PIN diode of type HPND-4005 of HP, which is beam lead type for high frequency operation. Capacitance value of the diode is 0.017 pF and resistance value varies from 2kΩ to 5Ω with diode current of 0 to 10 mA. The simulated amplitude and phase variation of the attenuator circuit with the change diode resistor is shown in Figure-4.1.5 [70]. Plot shows that PIN diode based attenuator provides very high level of attenuation up to 35 dB and also provides 180° phase shift, as required for vector modulator, when diode resistance crosses 50 Ω. Slight phase deviation near diode resistance of 50Ω is due to parasitic capacitance of of value 0.017 pF. The phase deviation can be eliminated by implementing inductance in parallel with the diode. The value of the inductance will be such that parallel resonance frequency will be the operating frequency of the vector modulator.

**Fig.-4.1.5**: Simulated $S_{21}$ plot of the variable attenuator
4.1.3 **Realization of Vector Modulator**

Schematic circuit diagram of the vector modulator circuit is shown in *Figure-4.1.6*. Here input signal at the input port is split equally between the direct and coupled ports, with a 90° phase difference. These two divided signals are selectively attenuated and phase shifted by the variable phase shifter and finally added by the Wilkinson power divider.

The amplitude $|S_{21}|$ and and phase $\angle S_{21}$ of the vector modulator is given by (4.1.4) and (4.1.5) respectively.

$$|S_{21}| = \left[ \frac{1}{2} \rho_1 \right]^2 + \left( \frac{1}{2} \rho_2 \right)^2 + \frac{1}{2} \rho_1 \rho_2 \cos \theta \right]^{1/2}$$  

(4.1.4)

$$\angle S_{21} = \tan^{-1} \frac{\rho_2 \sin \theta}{\rho_1 + \rho_2 \cos \theta}$$  

(4.1.5)
Where, \( \rho_1 \) and \( \rho_2 \) are the insertion loss of the attenuator-1 and attenuator-2 respectively and given by (4.1.2). And, \( \theta \) is the phase angle between the two signals combined at the output port of the vector modulator and given by (4.1.6)

\[
\theta = \phi_2 - \phi_1 - 90^\circ \quad (4.1.6)
\]

Where \( \phi_1 \) and \( \phi_2 \) are the insertion phase of the attenuator-1 and attenuator-2 respectively and given by (4.1.3).

In ideal case, minimum insertion loss is 3dB is due to the 90\(^\circ\) phase difference between the two signal combined at output. Simulated insertion loss is 4.5 dB. Simulated \( S_{21} \) of the vector modulator is shown in Figure-4.1.7. It shows that by varying diode resistances it is possible to achieve all phase and amplitudes.

Three different circuits are realized at three different frequency band of centre frequencies 2.6 GHz, 3.9 GHz and 11.6 GHz. The photograph of the circuits are shown in Figure-4.1.8. All the circuits are realized with MIC technology. RF circuits are realized on 25mil alumina (\( \varepsilon_r = 9.9 \)) subtrate. Beam lead PIN diode of type HPND4005 of HP is used for variable attenuator. The test results of the S-band vector modulator is

![Diagram](image_url)

**Fig.-4.1.7:** Simulated performance of the vector modulator circuit

\( C_d = 0.02 \) pF
Freq=2.6 GHz

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shown in Figure-4.1.9. The plot shows some sample of the readings. Here, PIN diodes are biased by constant voltage source by varying series resistor. Experimental achieved minimum insertion loss is 4.5 dB, when there is no bias to the attenuators. 7.5 dB attenuation is achieved when there is no bias to one attenuator and maximum attenuation is given to another attenuation. Constellation of vectors, of different amplitude and phase, as shown in the figure is achieved by giving different sets of bias to the two variable attenuators. This plot shows that all possible phase over 360° can be achieved with different resultant amplitude.

Fig.-4.1.8: Photograph of vector modulators

Fig.-4.1.9: Measured polar $S_{21}$ plot of S-band vector modulator
4.1.4 Temperature Behavior and Compensation of Vector Modulator

Variable attenuator of the vector modulator is realized using PIN diodes. RF resistance of PIN diode determines the amplitude and phase of the vector modulator. As discussed in the previous chapter, RF resistance of PIN diode is function of temperature. Thus, amplitude and phase of the vector modulator will also vary when the vector modulator will operate under varying temperature environment of spacecraft. Amount of variation depend upon the different amplitude and phase setting of the vector modulator. Worst-case amplitude and phase variation is nearly ±3dB and ±7.5° over the temperature range of -10 to +60 °C when the diodes are biased by conventional constant current bias condition. Constant current source bias not only vary the amplitude and phase performance of the vector modulator, but also it requires complicated circuit to realize variable constant current source to achieve constellation of amplitude and phase.

As discussed in the previous chapter, proposed optimum bias load line technology can provide temperature invariant attenuation of the PIN diode based attenuator. Same technology is applied to achieve temperature invariant amplitude and phase performance of the vector modulator.

**Fig.-4.1.10:** Schematic circuit of the temperature compensated analog vector modulator
Figure-4.1.10 shows circuits diagram of an analog vector modulator with temperature compensation biasing scheme [24]. Here, both the attenuators are biased from the same optimum voltage source \( V_{\text{OPT}} \) derived from single voltage source with the help of potential divider. The optimum voltage \( V_{\text{OPT}} \) is derived in previous chapter, which depends upon the PIN diode parameters and recalled here as:

\[
V_{\text{OPT}} = V_{\phi_0} + \frac{P}{2-m} (E_b - V_{\phi_0}) + \frac{\eta k T_0}{q} \left( \frac{pn}{2-m} - 1 \right) \quad (4.1.7)
\]

For PIN diode of type HPND4005, \( V_{\text{OPT}} = 1.2 \) volts. Variable resistors \( R_1 \) and \( R_Q \) are used to vary the currents of the attenuators continuously to vary amplitude and phase of the vector modulator continuously. With this optimum voltage source bias, amplitude and phase variation reduces to ±0.5dB and ±1.4° respectively over the temperature range of -10 to +60 °C. Thus, the proposed temperature compensated circuit provides stable constellation of amplitude and phase varying vector over very wide range of varying temperature environment.

For digital beamforming network, amplitude and phase of the vector modulator is to be changed digitally. Figure-4.1.11 shows the schematic circuit diagram of a digitally controlled temperature compensated vector modulator. Here electronic analog switch with bunch of selectable resistors are used to realize PIN diode based step attenuator. Depending upon data to the analog switch, only one resistor at a time will be connected to the input optimum voltage \( V_{\text{OPT}} \). The optimum voltage is determined by (4.1.7) by putting diode parameter, also can be determined graphically by collecting V-I data experimentally as discussed in the previous section. PROM is used to provide predetermined data to the analog switches to select any combination of resistors from both the switches, which will provide different combinations of current to the PIN diodes and produce different combination of temperature invariant amplitude and phase of the vector. Analog switch 74HC4051 is used to implement the digitally controlled vector modulator.
Fig.-4.1.11: Schematic circuit of the temperature compensated digitally controlled vector modulator

4.1.5 Conclusion

This section discussed a single circuit approach vector modulator using PIN diodes with novel optimum load-line biasing scheme for temperature invariant performance. Different circuits are realized in a balanced configuration with very good input and output return loss over different frequency of operation at 2.6 GHz, 3.9 GHz and 11.6 GHz. Due to its good input, output return losses, the circuit can be easily integrated with other circuits for beamforming network. The proposed circuit is very simple and provides temperature compensated amplitude and phase performance.
4.2 Temperature Compensated Diode Based Linearizer

Today's satellite communication transponder needs high power microwave amplifiers such as TWTAs and SSPAs to meet high EIRP requirements. Power amplifiers are require to operate at back off condition to meet required linearity during multi-carrier operation and high data rate digitally modulated traffic, which degrade DC to RF efficiency of the system. Linearizing is the technique used to improve non-linear performance of communication systems leads the high power amplifiers to operate at near saturation or at less back off condition. Thus, trade off between efficiency and linearity can be avoided by the use of linearizer.

Among the various types of linearizer [28] - [38], diode based pre-distortion linearizers are suitable for space applications for its compactness and less power consumption. There are numerous reported diode based pre-distortion linearizers [28] - [36] with their own merits and demerits. Pre-distortion linearizer generates amplitude and phase characteristic opposite to the characteristic of power amplifiers. Thus, cascading linearizer with power amplifier, decreases overall amplitude and phase non-linearity leads to overall improved linearity.

Fig.-4.2.1: Measured I-V characteristics of Schottky diode in presence of RF power
Measured I-V characteristic of a Schottky barrier diode (beam lead of type HSCH 5315) in presence of RF power is shown in Figure-4.2.1. Two types of load line, one for nearly constant voltage bias condition and another for nearly constant current bias condition are also shown in the plot. It is clear from the plot that RF resistance (dynamic resistance) of diode decreases with increase of RF power level when the diodes are biased with nearly constant voltage (small series resistance) bias condition. Whereas, RF resistance (dynamic resistance) of diode increases with increase of RF power level when diodes are biased with nearly constant current bias (high series resistance) condition. Depending on the load line effects, some linearizers are parallel diode based, biased by nearly constant current bias condition and some are series diode based, biased by nearly constant voltage bias condition. Some linearizers [36], [39], [40] are also uses PIN diode based attenuator with Schottky diode based distortion generator to generate adjustable amplitude and phase non-linearity characteristic to linearize various power amplifiers with different types of amplitude and phase non-linearity. Under conventional constant voltage or constant current bias condition, diode based linearizer’s performance changes with temperature since RF resistance of the diode is function of temperature as shown in

![Simulated RF resistance variation with temperature](image)

**Figure-6.2.2:** Simulated RF resistance variation with temperature

It shows that, series diode based linearizer, with nearly constant voltage bias condition will suffer severely with change of temperature compared to parallel diode based linearizer with nearly constant current bias condition.

In conventional approach, digital or analog circuits with temperature sensors can be used for temperature compensation of diode based linearizer, which lead to more part count and complex circuit, resulting less reliability. In previous chapters, we have shown...
that by selecting proper bias load line it is possible to achieve temperature invariant RF resistance of Schottky barrier diode and PIN diode.

This chapter discusses the design, development and test results of a novel diode based linearizer based on Schottky barrier diode and PIN diode. Temperature compensation of the proposed linearizer circuit, using novel diode biasing technique as discussed in the previous chapters, is also discussed.

### 4.2.1 Design of a Novel Diode Based Linearizer

Schematic circuit diagram of the proposed pre-distortion diode linearizer [39], [40] is shown in **Figure-4.2.3**. To achieve different combination of magnitude and phase characteristic, voltage variable PIN diode based (D1 and D2) attenuator circuit (labeled as linear arm) utilized in combination with distortion generator circuit (labeled as non-linear arm) using Schottky barrier diode (SD1 and SD2). These two arms are connected with a 3-dB 90° input power divider and a 3-dB in-phase output power combiner.

![Schematic of the proposed diode based linearizer](image)

**Fig.-4.2.3:** Schematic of the proposed diode based linearizer

Due to inherent property of 3-dB 90° hybrid at input and 3-dB in-phase output power combiner, input and output return loss will be always good for all values of the diode resistance as long as the diode pairs are matched. Thus, the proposed linearizer topology eliminates the use of isolators at both input and output ports, as required for the
previously reported many diode based linearizers [28] - [31]. Therefore, the proposed diode based linearizer leads to very compact linearizer circuit with very good input/output matching without using any isolators.

The equivalent circuits of a beam lead PIN diode and Schottky diode are consisting of an equivalent resistance $R_d$ with a parallel capacitance $C_d$ as shown in Figure-4.2.4. Therefore, insertion amplitude $\rho$ and phase $\phi$ of the 3-dB 90° coupler with two numbers of matched diode connected as shown in Figure-4.2.3, are given by equation (4.2.1) and (4.2.2) respectively.

\[ \rho = \left[ \frac{(R_d - R_o)^2 + (\omega C_d R_d R_o)^2}{(R_d + R_o)^2 + (\omega C_d R_d R_o)^2} \right]^{1/2} \quad (4.2.1) \]

\[ \phi = -\tan^{-1} \left[ \frac{2\omega C_d R_d^2 R_o}{R_d^2 - R_o^2 - (\omega C_d R_d R_o)^2} \right] \quad (4.2.2) \]

![Figure-4.2.4: RF equivalent circuit of forward biased junction diode](image)

Where $\rho = \rho_1$ and $\phi = \phi_1$ for $R_d = R_{pd}$ and $C_d = C_{pd}$ in case of coupler with PIN diode of linear arm.

And $\rho = \rho_2$ and $\phi = \phi_2$ for $R_d = R_{sd}$ and $C_d = C_{sd}$ in case of coupler with Schottky diode of non-linear arm.

The amplitude $|S21|$ and insertion phase $\angle S21$ of the linearizer is given by (4.2.3) and (4.2.4) respectively.

\[ |S21| = \left[ \left( \frac{1}{2} \rho_1 \right)^2 + \left( \frac{1}{2} \rho_2 \right)^2 + \frac{1}{2} \rho_1 \rho_2 \cos\theta \right]^{1/2} \quad (4.2.3) \]
\[ \angle S_{21} = \tan^{-1} \frac{\rho_2 \sin \theta}{\rho_1 + \rho_2 \cos \theta} \quad (4.2.4) \]

Where \( \theta \) is the phase angle between the two signals combined at the output port of the linearizer and given by (4.1.5).

\[ \theta = \phi_2 - \phi_1 - 90^\circ \quad (4.2.5) \]

**Figure-4.2.5** shows the plot of simulated amplitude and phase characteristic from equation (4.2.3) and (4.2.4). The plot shows the amplitude and phase of the linearizer with respect to Schottky diodes resistance for different setting of PIN diodes RF resistance at frequency of 11.55 GHz, taking \( C_{pd} = 0.01 \text{pF} \) and \( C_{sd} = 0.15 \text{pF} \). The plot shows that, the linearizer shows positive amplitude and negative phase deviation when diodes are biased by nearly constant current bias condition (i.e. RF resistance increases with increase of RF power level) and initial bias of the diodes are corresponds to RF resistance more than or equal to \( R_O \). This amplitude and phase characteristics (increase of amplitude and decrease of phase with increase of RF power level) is required to compensate the amplitude and phase non-linearity of solid state power amplifier (SSPAs) and traveling wave tube amplifiers (TWTAs). Other combinations of amplitude and phase characteristics can be obtained by different biasing conditions of the PIN and Schottky barrier diodes [39], [40].

A compact linearizer has been designed and developed [39] at the communication
satellite's Ku-band down-link frequency of 11.4 to 11.7 GHz. MIC layout of the complete linearizer circuit is realized very compactly of size 0.75 inch × 0.5 inch in a single 25 mil thick alumina substrate of ε_r = 9.9. The MIC assembly is shown in Figure-4.2.6. A 3-dB Lange coupler with two numbers of diode connected at the direct port and coupled port realizes linear and non-linear arms in a reflective type configuration. Isolated port of the coupler used as the output port of the linear and non-linear arm. For non-linear arm beam-lead Schottky barrier diode HSCH-5315 and for linear arm beam-lead PIN diode HPND-4005 has been used. Input power divider and output power combiner is realized by 3-dB Lange coupled and Willkinson power combiner respectively. R1 and R2 are the biasing resistors for liner and non-linear arms respectively. The complete circuit is analyzed using HPs Series-IV Harmonic-Balance simulator.

Putting bias resistor R2 of the non-linear arm before the high capacitance value line-filter LF2, minimizes the memory effect of the linearizer. Therefore, change of voltage and current of the bias resistor can follow the envelope of the multi-carrier signal. Thus, linearizer can be considered as a quasi-memory-less system [76], [77] and linearizer can be modeled only by their Pin/Pout characteristics (AM/AM conversion) and their input power dependent phase shift characteristics (AM/PM conversion). The photograph of the realized Ku-band linearizer is shown in Figure-4.2.7.
4.2.2 **Tested Results of a Ku-Band Linearizer**

Tested amplitude and phase performance of the Ku-Band linearizer is shown in Figure-4.2.8 at the frequency of 11.55 GHz. Here different combination of amplitude (AM/AM) and phase (AM/PM) characteristics is obtained for different bias conditions of the distortion generator and variable attenuator. Here $I_{dg}$ is the bias current of the distortion generator at no RF condition and $I_{att}$ is the bias current of the PIN diode attenuator. This figure shows that different combination of amplitude expansion up to 7dB and phase change up to 44 degree is achievable. Plot (A) for Schottky diodes current of 0.7 mA and PIN diodes current of 0.4 mA, generates gain expansion of 3 dB and phase decrease of nearly 20 deg. That is required to linearize SSPAs non-linearity. Plot (B) corresponds to, Schottky diodes current of 1.4 mA and PIN diodes current of 0.5 mA, generates gain expansion of 5 dB and phase decrease of nearly 35 deg. Whereas, plot (C) corresponds to Schottky diodes current of 1.5 mA and PIN diodes current of 0.6 mA, generates gain expansion of nearly 6.5 dB and phase decrease of nearly 46 degree. Both the combinations are suitable to linearize TWTAs to improve transmitter’s linearity, which uses high power tube amplifier.
4.2.3 **Evaluation of a S-Band Diode Linearizer**

An S-band diode based linearizer also realized [40] based on the same principle of operation as Ku-band linearizer discussed in the previous sections. The photograph of the realized S-band linearizer circuit is shown in Figure 4.2.9. The S-band linearizer designed at operating frequency of 2.55 to 2.63 GHz. 25 mil thick alumina (Al$_2$O$_3$) substrate of $\varepsilon_r = 9.9$ is used to realize all the MIC circuits. Linear and non-linear arms (to generate distortion) are realized by a 3-dB Lange coupler in a reflective type configuration. For non-linear arm beam-lead medium barrier batch matched Schottky barrier diode HSCH-5315 and for linear arm beam-lead PIN diode HPND-4005 has been used. Input power divider and output power combiner is realized by 3-dB Lange coupled and Willkinson power combiner respectively. All the individual circuits and integrated complete linearizer circuit is analyzed using HPs series IV Libra simulator (harmonic balance simulator).
All the MIC cards are brazed on the gold plated kovar carrier plates to minimize thermal stress of the alumina substrate and carrier plates are mounted in the aluminium box of size nearly 2 inch × 2 inch.

HPs Vector Network Analyzer (VNA) HP8510A is used for amplitude and phase characterization of the linearizer. The linearizer was evaluated with a 2-Watt S-band SSPA. The improvement of gain compression and total phase shift is shown in Figure-4.2.10. Gain compression decreases from 3.2 dB to 0.8 dB at nominal output power.

![Fig.-4.2.9: Photograph of S band linearizer](image)

![Fig.-4.2.10: Amplitude and phase characteristic of SSPA with linearizer.](image)
level. In addition, maximum total phase shift decreases from 6.2 degree to only 1 degree. *Figure-4.2.11* shows the improvement of third order inter-modulation (IM3) product at different output power back off. It shows that 3.8 dB IM3 improvement is at 1dB O/P back off and 8.7dB IM3 improvement is at 2 dB O/P back off.

![Fig-9: IMD of SSPA](image)

*Fig.-4.2.11: IMD of SSPA with and without linearizer*

### 4.2.4 Temperature Compensation

As discussed in the previous chapters that, RF resistance of Schottky barrier diode and PIN diode changes with temperature when diodes are biased by constant current or constant voltage bias conditions. Therefore, amplitude and phase characteristic of the linearizer will also vary with change of temperature and degrade the performance of the overall system.

*Figure-4.2.12* shows the measured amplitude variations and *Figure-4.2.13* shows the measured phase variation of the distortion generator circuit over different input power levels at three different temperatures of -10, +25 and +60 °C. Amplitude variation is nearly 5-dB at RF input level of -30 dBm and nearly 2-dB at -17 dBm. Whereas, drastic phase variation is observed at fixed current bias condition, over the entire RF power level, with the variation of diode temperature. Experimentally it is also observed, that to achieve temperature invariant performance of the distortion generator, requirement of
voltages/current is not only function of temperature but also it is function of input RF power level. Thus, for temperature compensation of the linearizer circuit for entire RF power levels, it is required to vary the voltage/current by sensing temperature as well as RF power level of the Schottky diode.

Here we have proposed a very simple technique [41], [21] - [24], which will ensure temperature invariant RF resistance of the individual diode leads to temperature invariant performance of the linearizer. Where for both PIN diode and Schottky barrier diodes are
biased by the optimum load line biasing scheme as discussed in chapter II.

*Figure-4.2.14* shows the measured amplitude and phase characteristic of the Schottky barrier diode based distortion generation circuit under optimum bias voltage of 0.75 Volts under different temperatures of -10, +30 and +60 °C. The measurement shows that practically there is negligible amplitude and phase variation with change of temperature at small signal condition. However, with the increase of RF power level, it shows slight amplitude and phase variation with change of temperature. At higher power level, this deviation is due to the self-bias effect of the Schottky diode. The measured plot shows that due to temperature change of -10 to +60 °C, over the RF power level of -30 to -10 dBm the amplitude variation remains within ±0.6 dB and phase variation remains within ±1.5 degree.

PIN diode based attenuator also separately tested with the proposed optimum load

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**Fig.-4.2.14:** Measured S21 variation with temperature for optimum load biasing

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PIN diode based attenuator also separately tested with the proposed optimum load biasing scheme [21] - [24] of optimum open circuit voltage of 1.19 Volts. The attenuation variation remains within ±0.2 dB over the temperature range of -10 to +60 °C.
Complete linearizer circuit is also tested over the temperature range of -10 to +60 °C. Measurement shows that the amplitude variation remains within ±0.5 dB and phase variation remains within ±1.2 degree over the entire operating RF power level of -30 to -10 dBm.

4.2.5 Conclusion

This chapter discussed a diode based novel linearizer for high power TWTAs and SSPAs. The proposed circuit is capable very wide range of amplitude and phase nonlinearity of TWTAs and SSPAs. Temperature behavior of the linearizer is also shown. Temperature compensation of the linearizer, using proposed optimum load line biasing technique shows that very simple bias circuit can provide temperature compensated RF performance over a very wide range of amplitude and phase variation. The compensation circuit is also applicable for wide range of RF power level over a broad temperature range.
4.3 A Temperature Compensated Over-Drive Level Controller For SSPAs

Reliability of GaAs power FETs used in SSPAs is affected adversely when operated under overdrive (excess input power) condition for a long time due to excess gate current. Manufacturer of SSPAs uses different schemes to protect FET power devices from over drive condition. One method for over-drive level control is by distributing compression in device line-up and another is by using limiter circuit. However, in both the cases it is very difficult to fully protect the devices from overdrive condition, which becomes worse when SSPAs are operated at varying temperature environment. Some manufacturer also uses feedback control loop for over drive control but its limiting power level is highly sensitive to temperature. Temperature compensated closed loop overdrive control circuit can eliminate all the above problems.

Here detailed description with mathematical expressions and test results of a temperature compensated closed loop overdrive level control (OLC) circuit is given [80]. The proposed scheme can properly protect FET power devices from over drive condition irrespective of change of temperature and given mathematical expressions helps to design the temperature compensation network without trial and error.

4.3.1 Different OLC Scheme

Manufacturer of SSPAs uses different types of schemes to protect GaAs power FET devices from over drive conditions.

In some cases, proper device line-up selected to distribute overdrive power among the devices in SSPA line up, so that power devices cannot be compressed heavily and gate current of GaAs FET devices remains within limit. However, in this case, device selection in the line up is not simple. Also precise overdrive protection of the device is
not possible without compromising the non-linearity of SSPAs, which becomes worse when SSPAs are operated at varying temperature environments.

Another approach to protect FET devices from overdrive power is use of Schottky barrier diode (SBD) based or GaAs FET based limiters [78], [79] at microwave frequency. However, the limiter can operate only at low power level range. Therefore, gain variation over temperature of the intermediate stages used in between power device and limiter circuit will change the limiting power level. Thus, in practice limiter circuits are also not suitable to properly protect the FET power devices used in SSPAs under over drive condition.

Some manufacturer of SSPAs uses closed loop power control systems in SSPAs for this purpose. However, due to high closed loop gain, the limiting power level is very much sensitive to variation of the detector diodes behavior over the temperature.

Temperature compensated closed loop over drive control circuit can eliminate all these disadvantages.

4.3.2 Operation of Closed Loop OLC

A simplified schematic block diagram of a closed loop overdrive level control system is shown in Figure-4.3.1. This scheme is capable to protect high power FET devices in power amplifier as well as small signal and medium power FET devices in driver

![Figure-4.3.1: Block diagram of proposed OLC](image)
amplifier from overdrive conditions. Since power sampler is used to sample and detect the power level, the scheme can be used to control any level of high power range. Photograph of the C-Band SSPA based on several small signal MESFETs (MGF1423) and power MESFETs (MGF2407 & MGF2430) of Mitsubishi is shown in Figure-4.3.2, which is used to demonstrate the proposed temperature compensated over-drive level control system.

Mainly Schottky diode detector is used at microwave frequency to detect microwave power. Control voltage $V_C$ is applied to the detector diode to finely adjust the limiting power level. When coupled power exceeds a certain specified level determined by the barrier potential ($\phi_B$) of Schottky barrier diode and given control bias $V_C$, then there will be detected output voltage. This detected voltage is amplified by DC amplifier and is fed to a voltage variable attenuator. Commonly PIN diode-based attenuators are used as a voltage variable attenuator. Depending upon required amount of over drive level protection, single or multi-stage attenuators can be used.

Test response of a closed loop OLC system is shown in Figure-4.3.3, where $P_{i,N}$ and $P_{o,N}$ are the normalized input and output power [80]. Here different limiting power level setting is achieved by controlling bias voltage ($V_C$) of the Schottky barrier detector diode.

![Figure-4.3.2: C-Band SSPA to demonstrate OLC Scheme](image-url)
Behavior of PIN diode attenuator and driver amplifier circuit of OLC system changes with temperature. RF resistance of PIN diode of variable attenuator circuit changes with temperature causing variation of attenuation of PIN diode attenuator. Gain of FET devices used in driver amplifier circuit also changes with temperature. Due to variation of attenuation of attenuator and gain of driver amplifier, the input power of the detector will vary with variation of temperature.

![Figure 4.3.3: Closed loop OLC response under different control bias voltages](image)

**4.3.3 Temperature Behavior of Closed Loop OLC**

*Fig.-4.3.4: I-V characteristics of SBD over temperature*
Also the \( I-V \) characteristic of detector diode changes with temperature as shown in the Figure-4.3.4. In case of hard limitation of power level, gain of DC amplifier is very high. Therefore, small change of detector diode voltage causes significant change of limiting power level. Tested results of variation of limiting power level of OLC system due to temperature variation of only the detector diode is shown in Figure-4.3.5. The limiting power level variation observed near about 3 dB peak to peak for the temperature variation of \(-10\) to \(+60\) °C.

### 4.3.4 Temperature Compensation of OLC System

Variation of limiting power level due to attenuation variation of attenuator and gain variation of driver amplifier are eliminated inherently due to the closed loop negative feedback behavior of the OLC system. Since gain of the DC amplifier is very high, when power level at the detector input changes due to the temperature variation then detector output changes automatically to adjust the attenuation of the attenuator to keep the input of the detector fixed.

![OLC response due to temperature variation](image)

**Fig.-4.3.5:** OLC response due to the variation of temperature of the detector diode

Therefore, as long as the detector diodes behavior is unchanged, power variation at the input of the detector, by any means, will be compensated by automatically changing the detected output voltage, which leads the change of attenuation of the attenuator such that power level at the input of the detector remains constant. Thus, attenuation and Gain
variation of the attenuator and FET devices of the driver amplifiers respectively will not affect the limiting power level of the OLC circuit if the voltage variation of the detector diode can be compensated over the temperature range of interest.

A temperature compensated OLC scheme is shown in Figure-4.3.6 [80]. The temperature dependent control voltage $V_C$ biases the detector diode $D_S$ through a bias network containing resistor $R_B$ and a RF decoupling network containing $\lambda/4$ transmission line and a capacitor $C_B$.

Output of the detector diode filtered by a low pass filter is then connected to the one input of the differential DC amplifier. Other input of the differential amplifier is connected to the control voltage $V_C$. Output of the differential amplifier is connected to the control terminal of the PIN diode attenuator circuit.

When RF power is below the limiting power level (threshold power level) then there will be no detector diode current, therefore, voltage at the two inputs of the differential amplifier will be same; it leads no output voltage and no attenuation of the attenuator. When the RF power amplitude crosses the threshold power level then detector diode starts to conduct. Therefore, there will be voltage difference between the inputs of the differential amplifier, which leads to a finite output voltage at the output of the
differential amplifier. This output voltage sets the attenuation of the attenuator so that the RF power will be limited to the predetermined limiting power level. Changing the control voltage \(V_c\) to the diode can compensate for the variation of limiting power level due to variation of detector diodes voltage \(V_{sbdl}\) over the temperature. For this, combination of detector diodes voltage and control voltage that is \((V_{sbdl} - V_c)\) should remain invariant over the temperature range of interest. Therefore, the condition for temperature compensation is:

\[
\frac{dV_c}{dT} = \frac{dV_{sbdl}}{dT} \quad (4.3.1)
\]

I-V characteristic over the temperature, for a p-n junction diode and Schottky barrier diode are similar in nature. Therefore, temperature dependent voltage source \(V_c\) generated by using temperature dependent voltage drop of a p-n junction diode D1.

In the circuit as shown in Figure-4.3.6, the control voltage \(V_c\) is given by:

\[
V_c = \frac{G}{1 + r} \left[ V_{pnd} - (1 - r)V \right] \quad (4.3.2)
\]

where, \(G = 1 + R_4/R_3\), \(r = R_2/R_1\) and \(V_{pnd}\) is the voltage of the p-n junction diode D1.

Therefore, combining equation (4.3.1) and (4.3.2), the condition for compensation becomes:

\[
\frac{G}{1 + r} \frac{dV_{pnd}}{dT} = \frac{dV_{sbdl}}{dT} \quad (4.3.3)
\]

Now, the temperature dependent I-V relationship of the diode is given by [1]:

\[
I_d = A T^n \text{Exp} \left( -\frac{qE_b}{\eta kT} \right) \left[ \text{Exp} \left( \frac{qV_d}{\eta kT} \right) - 1 \right] \quad (4.3.4)
\]

where, \(V_d\) is diode voltage, \(A\) is the area factor of the diode independent on temperature, \(\eta\) is diodes ideality factor, \(E_b\) is band gap potential of the semiconductor in case of p-n-junction diode and barrier potential \((\phi_b)\) in case of Schottky barrier diode, \(m\) is the temperature exponent of the diode, \(k\) is the Boltzman's constant and \(q\) is the electron charge.

And load line equation is given by:
\[ V_s = V_d + I_d \times R_b \quad (4.3.5) \]

where, \( V_d \) is the diode voltage at current \( I_d \) when biased by a voltage source \( V_s \) through a bias resistor \( R_b \). For forward biased diode where the reverse saturation current is negligible compared to the total forward current, the rate of change of diode voltage with respect to the temperature is given by [from expression (4.3.4) and (4.3.5)]:

\[
\frac{dV_d}{dT} = \frac{-I_d \cdot R_b}{(I_s \cdot R_b + \eta V_T)} \left[ \frac{E_b + m \eta V_T - V_d}{T} \right] \quad (4.3.6)
\]

where, \( V_T = k \cdot T / q \). This expression (4.3.6) is valid for both the p-n junction and Schottky barrier diode. Thus, knowing the diode parameters present in the expression (4.3.6), the circuit component values for the temperature dependent voltage source can be derived from expressions (4.3.2), (4.3.3) and (4.3.6).

### 4.3.5 Circuit Realization and Test Result

The schematic circuit as shown in Figure-4.3.6 has been implemented for a C-Band solid-state power amplifier. Voltage variable attenuator, using PIN diode HPND 4005, has been realized in alumina substrate (\( \varepsilon_r = 9.9 \)) by using 3-dB Lange coupler in a microstrip line configuration. Power sampler circuit and detector circuit using Schottky barrier diode HSCH 5315 also realized on the Alumina substrate. To generate the temperature dependent control voltage p-n junction diode 1N5806 used with supply voltage of ±5 Volts. High gain differential DC amplifier realized using operational amplifier (μA741).

Typical parameters for Schottky barrier diode HSCH-5315 are: \( E_b = 0.69 \) Volts, \( \eta = 1.1, m = 3 \) and \( V_{shd} = 0.28V \). Detector diode operates under very high bias resistor therefore, \( \eta \cdot V_T \ll I_d \cdot R_b \). Putting all these values \( dV_{shd}dT = 1.6 \) mV/°C at room temperature. And typical parameter for p-n junction diode 1N5806 are: \( E_b = 1.16 \) Volts, \( \eta = 1.5, m = 1.5 \) and \( V_{pnd} = 0.6 \) Volts. Therefore, \( dV_{pnd}dT = 2.1 \) mV/°C at room temperature. Putting these values to expression (6.2.2) and (6.2.3) and considering the p-n junction diode operates at current of 10 mA, and control voltage \( (V_c) \) requirement at room temperature is 0V, the circuit parameters becomes: \( R1 = 500\Omega, R2 = 440\Omega, R3 = 1k\Omega \) and \( R4 = 433\Omega \).
The test result of the temperature compensated OLC circuit is shown in Figure-4.3.7. From this figure, it is clear that limiting power level remains nearly invariant over the temperature range of -10 to +60 °C.

![Graph showing test result of temperature compensated OLC circuit](image)

**Fig.-4.3.7:** Complete OLC circuit response over the temperature variation

### 4.3.6 Conclusion

Here detailed description with mathematical expressions and test results of a temperature compensated closed loop overdrive level control (OLC) circuit is given. The proposed scheme can properly protect FET power devices from over drive condition irrespective of change of temperature. Thus, the circuit will be very much useful to protect sensitive MESFET, HFET, HFET microwave power devices from overdrive power conditions and will increase life of SSPAs. The given mathematical expressions help to design the temperature compensation network without trial and error.

Temperature compensated closed loop control circuit scheme presented here can be used not only for SSPAs but can be used in any microwave systems for power level...
control. These types of microwave power limiters are also suitable for signal processing for example to suppress AM components before demodulation of a microwave FM signal since PIN diode attenuator offers very low phase variation over the attenuation range.
4.4 A Temperature Compensated Channel Amplifier With Automatic Level Control For Satellite Transponder

Here a design, development and test results of a Ku-band channel amplifier with automatic level control (ALC) system for satellite transponder is discussed. A systematic temperature characterization and optimization procedure depending upon measurement data of the channel amplifier is presented. This procedure takes into account the effect of parameters variation from one unit to another and eliminates conventional trial and error method to determine the optimum component values of the temperature compensation circuits.

Satellite communication link at Ku-band frequency in tropical region faces excess up-link and down-link path loss due to rainfall. The up-link rain attenuation causes decrease in signal level received at satellite receiver which leads to decrease of satellite transmitter power. The signal level is further reduced by down link rain attenuation. This may cause signal level to fall below threshold level of the ground receiver for specific BER performance. Adaptive power control system at satellite transmitter [81], [82] and ground receiver can solve this problem. Here design, development and characterization of a Ku-band channel amplifier with automatic level control (ALC) system is discussed for spacecraft application to control input of the final power amplifier (TWTA or SSPA) according to the signal level arriving at the channel amplifiers input [83]. Thus, ALC system also protects the final power amplifier against any accidental high power from up-link as discussed in the previous section in over-drive-level control system.

The presented channel amplifier [83] can operate in ALC mode as well as in fixed gain mode (FGM). In fixed gain mode, gain will be 44 dB and in ALC mode, gain will vary automatically from 39 dB to 59 dB depending upon the input power level. This amplifier has a commandable gain control (22 dB) system to operate the final power amplifier in different back-off conditions in both modes of operation.

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The required gain of the amplifier is achieved by using three numbers of amplifier modules using HEMT devices (CFY67-08). In ALC mode, the channel amplifier operates as a closed loop feedback system. The amplifier contains a Schottky diode detector to detect the sampled RF power. The detected voltage amplified by a differential DC amplifier and applied to the control input of a variable PIN diode attenuator. Attenuation of the attenuator will vary according to the input power level so as to maintain the constant output power level of the channel amplifier. Another PIN diode attenuator is used for commandable step gain control.

It is known that gain of the HEMT based amplifier module, attenuation of the PIN diode attenuator and detected power level of the Schottky diode detector, all are function of temperature. Thus, suitable compensation circuits are included to compensate the temperature variation of the channel amplifiers performance over the qualification temperature range for satellite applications. A practical systematic procedure based on the measurement data instead of conventional trial and error method is presented to determine the component values of the compensation networks.

### 4.4.1 Block Schematic of the Channel Amplifier

The basic block schematic of the channel amplifier with ALC is shown in Figure-4.4.1. The amplifier module A1, A2 and A3 are used to meet the total gain requirement of the channel amplifier. The voltage variable attenuator AT1 is for automatic gain control of the amplifier depending on the input power level. The range of gain control by

![Block Schematic of the Channel Amplifier](image)

*Fig.-4.4.1: Basic block diagram of the channel amplifier with ALC*
this attenuator is +15 dB to -5 dB over the nominal gain of 44 dB. Attenuator AT2 is for
the commandable gain setting of the amplifier up to -22 dB, in steps of -2 dB.

For ALC function, RF power is sampled after amplifier A2 and detected by the
detector diode. This detected voltage is applied to one of the input of the differential DC
amplifier. Other input of the differential amplifier is connected to the reference voltage
$V_{R2}$. Voltage $V_{R2}$ will determine the output power level of the channel amplifier. Output
of the DC amplifier connected to the ALC attenuator AT1, through an analog switch.
Switch position will determine the mode of operation of the channel amplifier. When the
switch connects attenuator AT1 to the other reference voltage $V_{R1}$, then the operation of
the channel amplifier will be in fixed gain mode (FGM). Reference voltage $V_{R1}$ will
determine the gain of the channel amplifier in fixed gain mode operation.

The gain of the microwave amplifier modules, attenuation of the PIN diode
attenuators and detected power level of the Schottky diode detector are all functions of
the temperature. Thus, suitable compensation circuits are required to compensate the
temperature variation of the channel amplifiers performance over the qualification
temperature range for satellite transponder.

The schematic circuit diagram of the different temperature dependent control signals
is shown in Figure-4.4.2. Here, temperature controlled reference voltage $V_{R1}$ is generated
by using thermistor to compensate the gain in FGM operation. Whereas, temperature
controlled reference voltage $V_{R2}$ is generated by using P-N junction diode to compensate
the output power level variation in ALC mode operation. To achieve temperature
invariant accurate step attenuation, proposed optimum bias load line technique used with
analog switch CD4051 and bank of resistors, as discussed in chapter-II for PIN diode
attenuator.

4.4.2 Realization of Individual Modules

Amplifier modules A1, A2 and A3 are used to achieve required channel amplifier
gain. A1 & A2 are the three-stage amplifier and A3 is a two-stage amplifier. All the
amplifier modules are realized using pHEMT device of type CFY67-08 for Ku-Band
Fig.-4.4.2: Block diagram with control signals.
downlink frequency of 11.45 to 11.7 GHz. MIC assembly drawing of the three stage amplifier is shown in Figure-4.4.3.

Reactive matching networks are used for input and output matching of the devices. Resistive loading with open circuit stubs are used at input and output cards to achieve high stable amplifier. Drain and gate bias resistors with appropriate capacitors are accommodated within the MIC cards, which leads to very compact circuit with improved out-of-band stability.

![MIC assembly of 3-stage amp.](image)

*Fig.-4.4.3: MIC assembly of 3-stage amp.*

![MIC assembly of PIN diode attenuator](image)

*Fig.-4.4.4: MIC assembly of PIN diode attenuator*

All the matching networks of amplifiers are realized in a 25 mil alumina (Al₂O₃) substrate ($\varepsilon_r = 9.9$) of size $0.25'' \times 0.5''$, with the accommodation of all the bias resistors in the alumina substrate at the RF tray. Two-stage amplifier A3 is realized with the same MIC cards used in the three-stage amplifier, by eliminating one inter-stage matching.
network of three-stage amplifier. These circuits are simulated and analyzed by series IV circuit simulator of HP (EEsof).

To adjust overall gain of the channel amplifier the resistors $R_{D1}$, $R_{D2}$ & $R_{D3}$ are used in the +V1 supply lines of the amplifier A1, A2 & A3 respectively.

Attenuator for step gain control (22 dB) and automatic level control (20 dB) are both the two-stage PIN diode based voltage variable analog attenuator. MIC assembly drawing of the attenuator is shown in Figure -4.4.4. D1, D2, D3 & D4 are the beam lead PIN diodes (MPND-4005) mounted in a 3-dB Lange coupler designed at Ku-band frequency. In this configuration, the circuit will provide maximum attenuation when PIN diode resistance will be 50Ω which will be determined by current through the diode.

The photograph of the integrated RF tray is shown in Figure-4.4.5.

**Fig.-4.4.5:** Ku-band channel amplifier with ALC Scheme
4.4.3 Temperature Compensation of the Channel Amplifier

It is known that gain of HEMT based amplifier module, attenuation of PIN diode attenuator and detected power level of Schottky diode detector are function of temperature. Moreover, temperature dependency of on resistance of analog switch (CD4051) is influence the temperature characteristic of the channel amplifier. Thus suitable temperature compensation circuits, with suitable method to determine the component values, are required to compensate the temperature variation of the channel amplifiers performance over the qualification temperature range for satellite applications. There are various temperature compensation schemes for ALC loop compensations [30], however, here a novel techniques are used for this purpose [83]. There are two temperature- dependent control signals $V_{R1}$ and $V_{R2}$ generated to compensate the temperature variation of the channel amplifiers gain at FGM and ALC mode respectively. Another control signal $V_{R3}$ (Figure-4.4.2) is generated to achieve temperature invariant step gain setting. The practical procedures to determine the component values of the control circuits in different mode of operations are discussed in the following sections.

4.4.3.1 Temperature Compensation for Fixed Gain Mode

In fixed gain mode it is to be ensured that, gain of the channel amplifier will be within the specified limit over the operating temperature range. The function of the control signal $V_{R1}$ is (a) to set the gain of the amplifier 44 dB and (b) this gain should remain within the specified limit (peak to peak <0.8 dB) over the temperature range of -10 to +60 °C.

To generate temperature controlled reference signal $V_{R1}$, thermistor $R_{TH}$ used with the resistors $R_{F1}$, $R_{F2}$ & $R_{F3}$ as shown in Figure-4.4.2.

Equivalent circuit of the $V_{R1}$ signal generator with attenuator and analog switch is shown in Figure-4.4.6. Here, reference point taken before analog switch to include its (resistance of analog switch) temperature variation. Where $V_{PO}(T)$ and $R_{PO}(T)$ are the
function of temperature due to dependency on $R_{TH}(T)$ and given by the equations (4.4.1) and (4.4.2) respectively.

$$V_{FO}(T) = \frac{V_1[R_{F2} \parallel (R_{F3} + R_{TH}(T))] - R_{F1} [R_{F2} \parallel (R_{F3} + R_{TH}(T))]}{R_{F1} + [R_{F2} \parallel (R_{F3} + R_{TH}(T))]} \quad (4.4.1)$$

$$R_{FO}(T) = R_{F1} [R_{F2} \parallel (R_{F3} + R_{TH}(T))] \quad (4.4.2)$$

Suppose at temperature $T$, voltage at the reference point and current to the attenuator (as shown in Figure-4.4.2) is $V_{Ri}(T)$ and $I_{Ri}(T)$ to achieve required gain (44 dB) setting of the channel amplifier, then from the circuit of Figure-4.4.6 we can write the equation (4.4.3).

$$V_{FO}(T) - V_{Ri}(T) = R_{FO}(T) \times I_{Ri}(T) \quad (4.4.3)$$

To determine the circuit component values ($R_{F1}, R_{F2} \& R_{F3}$) of this network it is required to take three sets of voltage and current readings $\{V_{Ri}(T_a), I_{Ri}(T_a)\}; \{V_{Ri}(T_c), I_{Ri}(T_c)\}; \{V_{Ri}(T_h), I_{Ri}(T_h)\}$ at three different temperatures ($T_a, T_c, T_h$) by setting the amplifiers gain at the required value (44 dB). The optimum value of $R_{F1}, R_{F2} \& R_{F3}$ will be the one which satisfies the following three equations (6.4.4A), (6.4.4B) & (4.4.4C) simultaneously.

$$V_{FO}(T_a) - V_{Ri}(T_a) = R_{FO}(T_a) \times I_{Ri}(T_a) \quad (4.4.4A)$$
$$V_{FO}(T_c) - V_{Ri}(T_c) = R_{FO}(T_c) \times I_{Ri}(T_c) \quad (4.4.4B)$$
$$V_{FO}(T_h) - V_{Ri}(T_h) = R_{FO}(T_h) \times I_{Ri}(T_h) \quad (4.4.4C)$$

**Fig.-6.4.6:** Equivalent circuit for $V_{Ri}$
Based on these three equations the characteristic of the temperature-compensation circuit, in fixed gain mode, is modeled using a spreadsheet. The three unknown parameters \( R_{F1} \), \( R_{F2} \) and \( R_{F3} \) were varied interactively until all these three equations (4.4.4A, B, C) satisfied simultaneously. The particular values of \( R_{F1} \), \( R_{F2} \) and \( R_{F3} \) for which all these three equations satisfied simultaneously with minimum error, will be the optimum value of these resistors to provide temperature compensated gain in fixed gain mode operation of the channel amplifier.

For utmost accuracy the resistance of the actual thermistor, ultimately to be used in the compensation circuit, should also be recorded at three different temperatures. The three temperatures should be stabilized before taking the measurements to minimize the thermal gradient between thermistor and the circuit components.

### 4.4.3.2 Temperature Compensation for ALC Mode

In ALC mode, the detector circuit will detect RF power, and there will be voltage drop across the load resistor \( R_{A10} \). This voltage will be amplified by the differential amplifier and fed to the attenuator AT1 through analog switch. The reference voltage \( V_{R2} \) will determine output of the channel amplifier in ALC mode. Variation of the output power over the range of the input power variation will be determined by the overall gain of the ALC loop. Loop gain is determined by the resistor \( R_{A8}, R_{A9}, R_{A10} \ & R_{A11} \). Offset current provides through the resistor \( R_{A7} \) will ensure the output of the differential amplifier at negative potential at very low RF power level.

Characteristic of the detector, amplifier modules and attenuators are changes with temperature. Thus, reference voltage \( V_{R2} \) should be a temperature dependent voltage to achieve required fixed output power level within the specified range over the operating temperature range.

To generate temperature controlled reference voltage \( V_{R3} \), p-n junction diodes are used as shown in the Figure-4.4.2. The equation (4.4.5) and (4.4.6) will determine current \( (I_d) \) through the diodes and voltage \( (V_d) \) across the all diodes.
\[ I_d(T) = f(V_d(T)) \quad (4.4.5) \]

\[ I_d(T) = \frac{V_1 - V_B(T) - I_d(T) \times R_{A2} + V_2 - \frac{V_d(T) + I_d(T) \times R_{A2} - V_2}{R_{A3} + R_{A4}}} \quad (4.4.6) \]

\[ V_{R2}(T) \times [R_{A3} + R_{A4}] = [V_d(T) + I_d(T) \times R_{A2} - V_2] \times R_{A4} \quad (4.4.7) \]

We can assume that there is no current drawn by the detector diode & differential amplifier from the reference supply \( V_{R2} \). Therefore, at a particular temperature \( T \), if the reference voltage requirement is \( V_{R2}(T) \) then from Figure-4.4.2 one can write equation (4.4.7).

The circuit component values \( (R_{A1}, R_{A2}, R_{A3} \text{ & } R_{A4}) \) of this network can be determined by taking three sets of voltage readings \( [V_{R2}(T_A), V_{R2}(T_C), V_{R2}(T_H)] \) at three different temperatures \( (T_A, T_C, T_H) \) by setting the amplifiers output power level at the required value (say 0-dBm). Then the optimum value of \( R_{A1}, R_{A2}, R_{A3} \text{ & } R_{A4} \) will be the one which satisfies the following three equations (4.4.8P), (4.4.8Q), and (4.4.8R) simultaneously.

\[ V_{R2}(T_A) \times [R_{A3} + R_{A4}] = [V_d(T_A) + I_d(T_A) \times R_{A2} - V_2] \times R_{A4} \quad (4.4.8P) \]

\[ V_{R2}(T_C) \times [R_{A3} + R_{A4}] = [V_d(T_C) + I_d(T_C) \times R_{A2} - V_2] \times R_{A4} \quad (4.4.8Q) \]

\[ V_{R2}(T_H) \times [R_{A3} + R_{A4}] = [V_d(T_H) + I_d(T_H) \times R_{A2} - V_2] \times R_{A4} \quad (4.4.8R) \]

Based on these equations 4.4.5, 4.4.6 and 4.4.8P, Q, R, the characteristic of the temperature-compensation circuit, in ALC mode, are modeled using a spreadsheet. The four unknown parameters \( R_{A1}, R_{A2}, R_{A3} \text{ and } R_{A4} \) were varied interactively until all these three equations (4.4.8P, Q, R) satisfied simultaneously. The particular values of \( R_{A1}, R_{A2}, R_{A3}, \text{ and } R_{A4} \) for which all these three equations satisfied simultaneously with minimum error, will be the optimum value of these resistors to provide temperature compensated output power level in ALC mode operation of the channel amplifier.
Step attenuator AT2 is for the commandable gain setting of the channel amplifier. Proper load line selection technique [21] - [24], [83] is used to achieve temperature invariant attenuation of the PIN diode based attenuator AT2. Analog switch (CD4051) with a bank of resistors $R_{SI}$ to $R_{SN}$ and reference voltage $V_{R3}$ is used for the control signal of the PIN diode attenuator circuit.

Reference voltage $V_{R3}$ is the critical parameter to achieve temperature invariant attenuation. The resistor $R_{R1}$ and $R_{R2}$ determine this reference voltage. To determine $R_{R1}$ and $R_{R2}$, it is required to take three sets of voltage and current readings $\{V_{AT2}(T_{A}), I_{AT2}(T_{A}), V_{SW}(T_{A})\}; \{V_{AT2}(T_{C}), I_{AT2}(T_{C}), V_{SW}(T_{C})\}; \{V_{AT2}(T_{H}), I_{AT2}(T_{H}), V_{SW}(T_{H})\}$ at three different temperature $(T_{A}, T_{C}, T_{H})$ by setting the attenuation of the attenuator preferably at the maximum required value (say 22-dB). Then plotting these data, as shown in Figure-4.4.7, one can get reference voltage $V_{R3}$. Where, $V_{R3}$ is the intercept point of the load line with the voltage $(V_{AT2}+V_{SW})$ axis. Then $R_{R1}$ and $R_{R2}$ will be determined by satisfying the equation (4.4.9) and putting the constraint on current drawn from the source $V_{I}$.

$$V_{R3} = \frac{V_{I} \times R_{R2}}{R_{R1} + R_{R2}} \quad (4.4.9)$$

The resistance values $R_{SI}$ to $R_{SW}$ can be determined by setting the reference voltage $V_{R3}$ and then adjusting these resistor values to achieve different required step attenuation (2, 4, 6,--- dB). This can be done at any temperature.

![Figure-4.4.7: Plot to determine $V_{R3}$](image-url)
In this method, temperature variation of the 'analog switch resistance' has been taken into account by plotting the load line with respect to voltage \((V_{AT2} + V_{SW})\). Therefore, this procedure eliminates the separate temperature characterization of the 'analog switch resistance' over the temperature.

### 4.4.4 Measurement and Test Results

The integrated channel amplifier has been fabricated and tested over the Ku-band down-link frequency of 11.45 to 11.7 GHz. Measured I-O characteristics of the channel amplifier in ALC mode, without temperature compensation, is shown in Figure-4.4.8. Measured temperature compensated I-O characteristic in ALC mode is shown in Figure-4.4.9. The measured output power variation is within 1-dB over the input power variation of -59 to -39 dB.

As shown in Figure-4.4.8, without temperature compensation circuit, the output power variation in the ALC mode over the temperature range of -10 °C (cold) to +60 °C (hot) is nearly 4.5 dB. The required reference voltage \((V_{R2})\) is measured at cold, ambient

![Fig.-4.4.8: I-O characteristic without temperature compensation.](image)

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equations (4.4.8P), (4.4.8Q) & (4.4.8R) simultaneously. I-O characteristic of the channel amplifier with the optimum components value of temperature compensation circuit in the ALC mode is shown in Figure-4.4.9 at three different temperatures. With temperature-compensation circuit, the measured variation reduces to less than 0.4 dB over the temperature range of -10 to +60 °C.

In fixed gain mode, without compensation circuit the gain variation of the channel amplifier is nearly 5-dB over the temperature range of -10 to +60 °C. The three sets of voltage and current readings \(V_{R1}, I_{R1}\) at cold, ambient & hot temperatures are measured to get same gain of 44 dB. The optimum values of the resistors \(R_{F1}, R_{F2}, R_{F3}\) of the compensation network are determined for 5 kΩ thermistor by satisfying equations (4.4.4A), (4.4.4B) & (4.4.4C) simultaneously. With the optimum components value of the compensation network, the measured gain variation over the temperature range becomes 0.4 dB p-p.

The optimum reference voltage \(V_{R3}\) is determined by the optimum load line selection technique. The three sets of voltage and current \([(V_{A72} + V_5), (I_{A72})]\ readings at cold, ambient & hot temperatures are measured to get same attenuation of 22 dB. Plotting these data, the optimum voltage \(V_{R3}\) is 0.963 Volts. For this optimum reference voltage,
the resistor values $R_{ri}$, $R_{r2}$, and $R_{si}$ to $R_{sn}$ are determined. With this reference voltage, the achieved step attenuation accuracy is within ±0.4 dB for all the steps (-2, -4, ..., -22 dB) over the temperature range of -10 to +60 °C.

### 4.4.5 Conclusion

This paper discusses the design, temperature characterization and test results of a Ku-band channel amplifier with automatic level control (ALC) system for spacecraft application. The presented practical procedure to determine the component values of the compensation circuits will be very much useful to optimize the performance of the channel amplifier without characterizing the individual modules. This systematic procedure takes into account the effect of parameter variation from one unit to another and eliminates conventional trial and error method leads to minimize man hour time.
4.5 CONCLUSION

This chapter discussed the temperature behavior and compensation mechanism of different microwave subsystems. Mathematical and practical design procedures are also developed and discussed to determine the circuit component values. Different devices are temperature compensated by different techniques depending upon the function of the devices. For example, in case of PIN diode and Schottky barrier diode, optimum load line technique is used in case of vector modulator and linearizer application, in which, temperature insensitive RF resistance is required. Whereas, P-N junction diode is used for OLC and ALC application, in which it is used as RF detector. Overall gain of the channel amplifier based on MESFET and HEMTs are stabilized using temperature dependent variable PIN diode based attenuators. In all the cases temperature invariant attenuation is achieved with the use of PIN diode attenuator with proposed optimum load line biasing scheme.