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FILTER SYNTHESIS TECHNIQUES

Georgia Institute of Technology

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FILTER SYNTHESIS TECHNIQUES

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FOREWORD

This report was prepared by the Electronics Division of the Georgia Institute of Technology, Hinman Research Building, Atlanta, Georgia, under Contract F30602-68-C-0080, Project Number 4540, Task Number 454003. Secondary report number is A-1058-F. The work described was performed under the general supervision of D. W. Robertson, Head, Communications Branch, and H. W. Denny, Project Director. The authors are H. W. Denny and Charles S. Wilson. The contributions of E. E. Donaldson, Jr. and R. A. Byers to the project are acknowledged. The RADC Project Engineer was George A. Long (EMNCI-2).

This technical report has been reviewed by the Office of Information (EMLS) and is releasable to the Clearinghouse for Federal Scientific and Technical Information.

This technical report has been reviewed and is approved.

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ABSTRACT

Several active filter techniques for the reduction of receiver interference in the 225 to 400 MHz range are described. Positive feedback Q multiplier techniques were extended to include (1) the use of multiple feedback loops to achieve a high order of stable multiplication in each stage and (2) the use of cascaded stages of Q-multiplied resonators to obtain improved skirt selectivity. Negative resistance Q multiplication was achieved over a wide frequency range through the development of a common collector transistor amplifier that exhibits stable negative resistance properties in the UHF region. The negative resistance amplifier was incorporated into a breadboard model of a tunable filter which employs both active and passive stages to produce a high Q response characteristic with high skirt selectivity over the entire band. An AM cancellation filter that achieves suppression of an unwanted signal by cancellation via a synthesized replica of the signal was developed. The breadboard model demonstrated a suppression capability of 30-35 dB for AM signals and about 50 dB for CW signals.

To enhance the capabilities and versatility of UHF active interference suppression filters, linearization techniques for broadband solid state amplifiers were investigated. The application of negative feedback and the use of the push-pull mode of parallel operation provided a significant reduction in harmonic generation while retaining good gain-bandwidth characteristics in amplifiers of one-watt power output capabilities.

EVALUATION

The objective of this work was to discover new methods of designing active filter circuits that would allow closer frequency spacing in collocated communication systems. The effort was concentrated in the 225 to 400 MHz range where the problem was most prevalent. Three avenues of effort were explored in this contract:

1. The development of a three-stage cascaded UHF Q Multiplier.

2. The development of a UHF AM Cancellation Filter.

3. Research into techniques to improve the linearity of broadband UHF amplifiers.

In the development of the Q Multiplier the objective of successively multiplying the Q of three cascaded cavities was not met. Only one cavity actually provided multiplication. The other two were only optimumly coupled without multiplication. Also this multiplication was sufficiently effective over a frequency range of about 20 MHz at the upper end of the UHF frequency allocation, that is from 370 to 390 MHz. There are many problems of cavity tracking and multiple feedback adjustments which must be solved to produce a highly selective multi stage Q multiplier. With the discovery of better voltage controlled tuning devices and high quality stripline resonators development of a multi stage Q multiplier with superior filtering characteristics may become a practicality. Further exploration into this technique could produce worthwhile results. I would say that the objective of this phase was partly accomplished. The difficulties already mentioned prevented total accomplishment.

In the UHF AM Cancellation Filter Development, the feasibility of synthesizing a cancellation signal was proved and the objective accomplished. However, the process is excessively complicated. With the development of more intricate integrated circuits, realistic devices using this principle could be used.

Some work was done on the reduction of harmonics and intermodulation products of broadband transistorized amplifier by linearity improvement. The overall results of this work produced amplifiers that are equivalent to, but no better than, what is commercially available. The objective was not accomplished here. This is an extremely difficult undertaking. The problem lies more in the characteristics of the transistor than in the external circuitry. Significant improvement in transistor characteristics may be necessary before satisfactory linear amplifiers are produced.

GEORGE A. LONG
Effort Engineer

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SECTION I

INTRODUCTION

Increased demands for communications channels in the 225 to 400 MHz range have made it necessary to reduce the widths of the guardbands used to separate these channels. The resulting spectrum crowding has greatly increased the problem of adjacent and co-channel interference. The use of higher powered transmitters and more sensitive receivers has intensified the problem. This combination of high level signals, highly sensitive receivers, and close channel spacing places severe requirements on RF filters for interference reduction.

Crowding of the spectrum between 225 and 400 MHz is approaching the point where conventional passive devices do not provide adequate filtering action. Passive devices are limited in the degree to which they can solve current interference problems because, to provide the required Q and skirt selectivity, they must be impractically large.

In recent years, the technology of the design and analysis of active filters has progressed rapidly. The initial impetus to active filter development has been provided by low frequency requirements. The use of active techniques in the low frequency region has made it possible to construct physically small filters which are compatible with miniaturized systems.

In view of the advantages which have been demonstrated at lower frequencies, i.e., improved performance and/or significant reduction in the size of filters, a study of the application of active filter techniques in the VHF and UHF regions was performed at Georgia Tech under contract F30602-C-67-0066 [1]. Through the efforts of this program and a previous program [2], the feasibility of signal cancellation and Q multiplication in the 225 to 400 MHz range has been verified.

Although the active filters demonstrated the feasibility of specific techniques, they were limited in their power handling capabilities and/or in their frequency range of operation. In particular, the Q multiplication technique was limited by the power handling capabilities of the available broadband solid state amplifiers. Consequently, a major emphasis in this program was placed on the development of higher powered amplifiers with a high degree of linearity.

The further development of Q multiplication techniques was pursued through the use of multiple feedback loops to stably obtain high orders of multiplication, the use of cascaded stages of multiplied sections to obtain greater skirt selectivity, and the application of negative resistance techniques to obtain a greater operating frequency range.

Negative resistance Q multiplication was employed in a breadboard model of a tunable bandpass filter to produce an improved Q response with high skirt selectivity over the entire 225 to 400 MHz range.

This report describes the activities in the above areas and summarizes the findings of these studies. In addition, a description of a refined AM cancellation system is presented to demonstrate the suppression capabilities of a technique which synthesizes the interfering signal and uses the synthesized replica to cancel the unwanted signal.

SECTION II

Q MULTIPLICATION TECHNIQUES

A. Positive Feedback

The basic positive feedback Q multiplier configuration is shown in Figure 1. To realize very high Q response characteristics with a single loop, either a resonator with a high initial Q must be employed or a very high multiplication factor must be realized. High loaded Q's in passive devices are generally achieved only in physically large devices. Such large devices are incompatible with compact solid state equipments and, consequently, the smallest sized resonator is desired. Unfortunately, the smaller devices exhibit lower Q's which require a high multiplication factor. System stability becomes difficult to maintain as the multiplication factor increases [3].

One method that can be used to avoid the instability problems normally associated with very high multiplication factors is the use of successive stages of multiplication. Figure 2 illustrates the use of multiple stages to obtain a high effective multiplication factor. By achieving the desired multiplication factor in two steps rather than in a single step, improved stability can be expected. For example, a small coaxial cavity with a natural Q of 160 was incorporated into a two stage multiplier to achieve an effective Q greater than 4000. The net multiplication factor was 25 yet the system stability was comparable to the stability of a single stage multiplier with a multiplication factor of 6 to 7. It is evident that successive stages of Q multiplication can be used to realize extremely narrow bandwidth responses from passive resonators of moderate Q.

Although the 3 dB bandwidth is commonly used as a figure of merit for comparing various resonant structures, other characteristics are important for the reduction of adjacent channel interference. For example, the skirt selectivity is a measure of the rate of increase of attenuation outside the immediate passband. The selectivity of a bandpass filter is a function of the number as well as the Q of resonators in the filter.

The substitution of multiple resonators for a single resonator in the Q multiplication network immediately suggests itself as a means for achieving high Q and steep skirt selectivity. However, this approach is difficult to realize [4] because the phase shift through the resonators is difficult to match in the feedback loop. Figure 3 illustrates a possible approach to obtaining high Q performance with increased skirt selectivity. In this system, two active filters are cascaded which effectively doubles the attenuation roll-off rate on the skirts of the selectivity curve without degrading the Q of the individual filters. This system has the advantage that either the filters may be synchronously

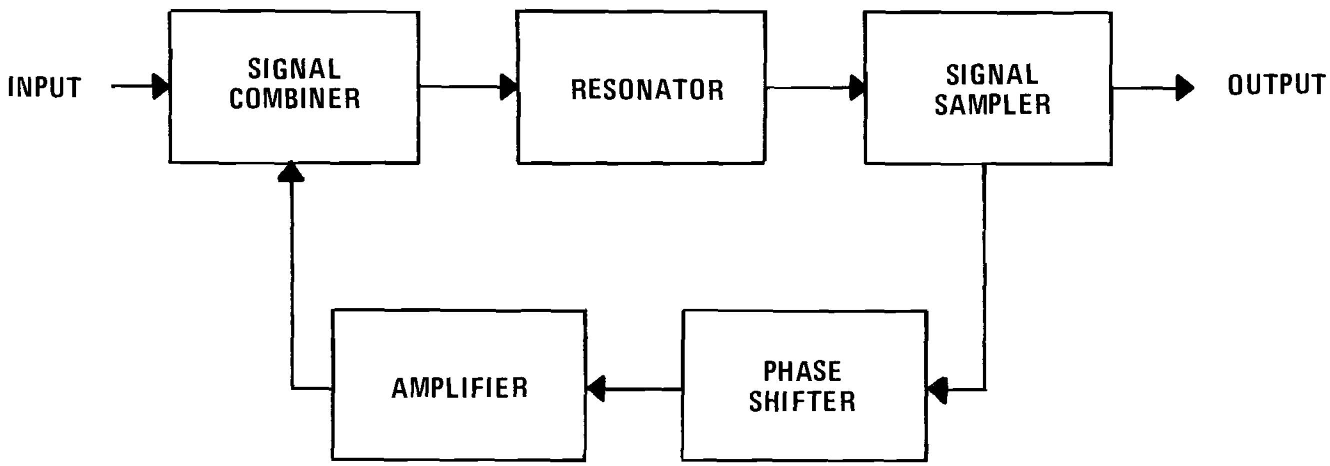


Figure 1. Block Diagram of the Basic Q Multiplication Network Configuration.

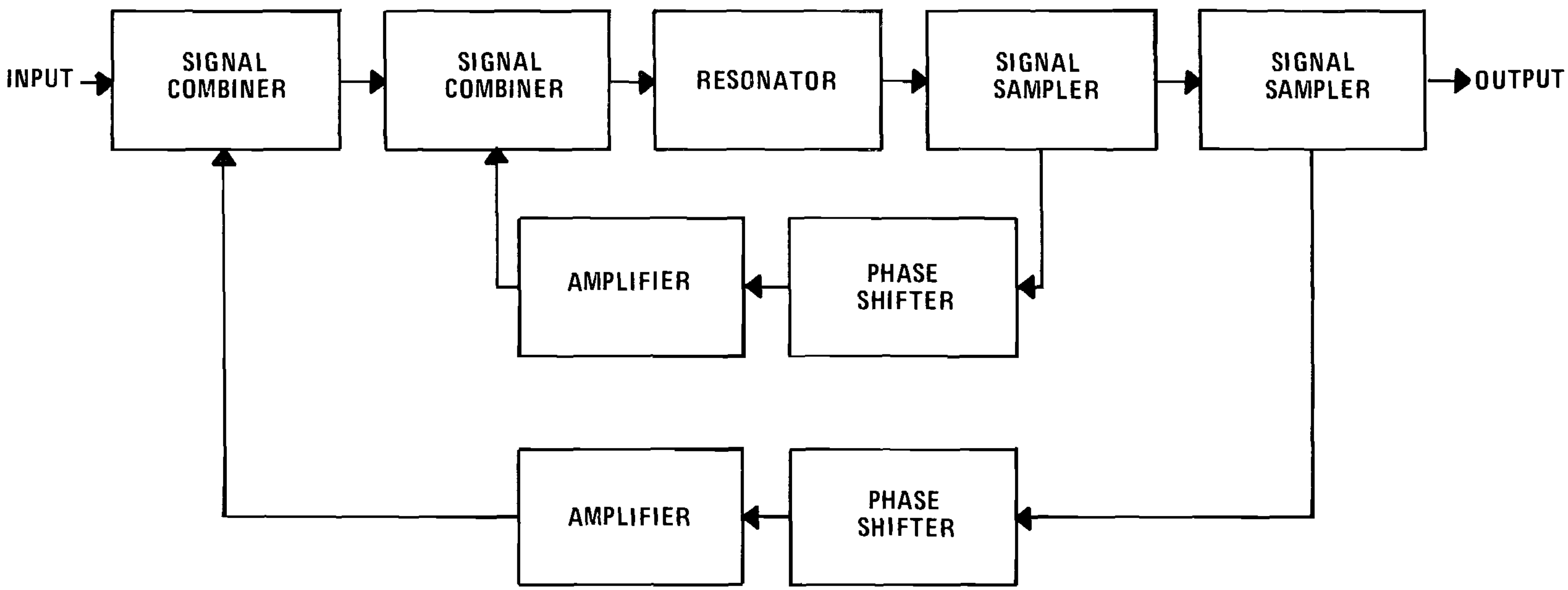


Figure 2. Block Diagram Showing the Two Stage Q Multiplication Technique.

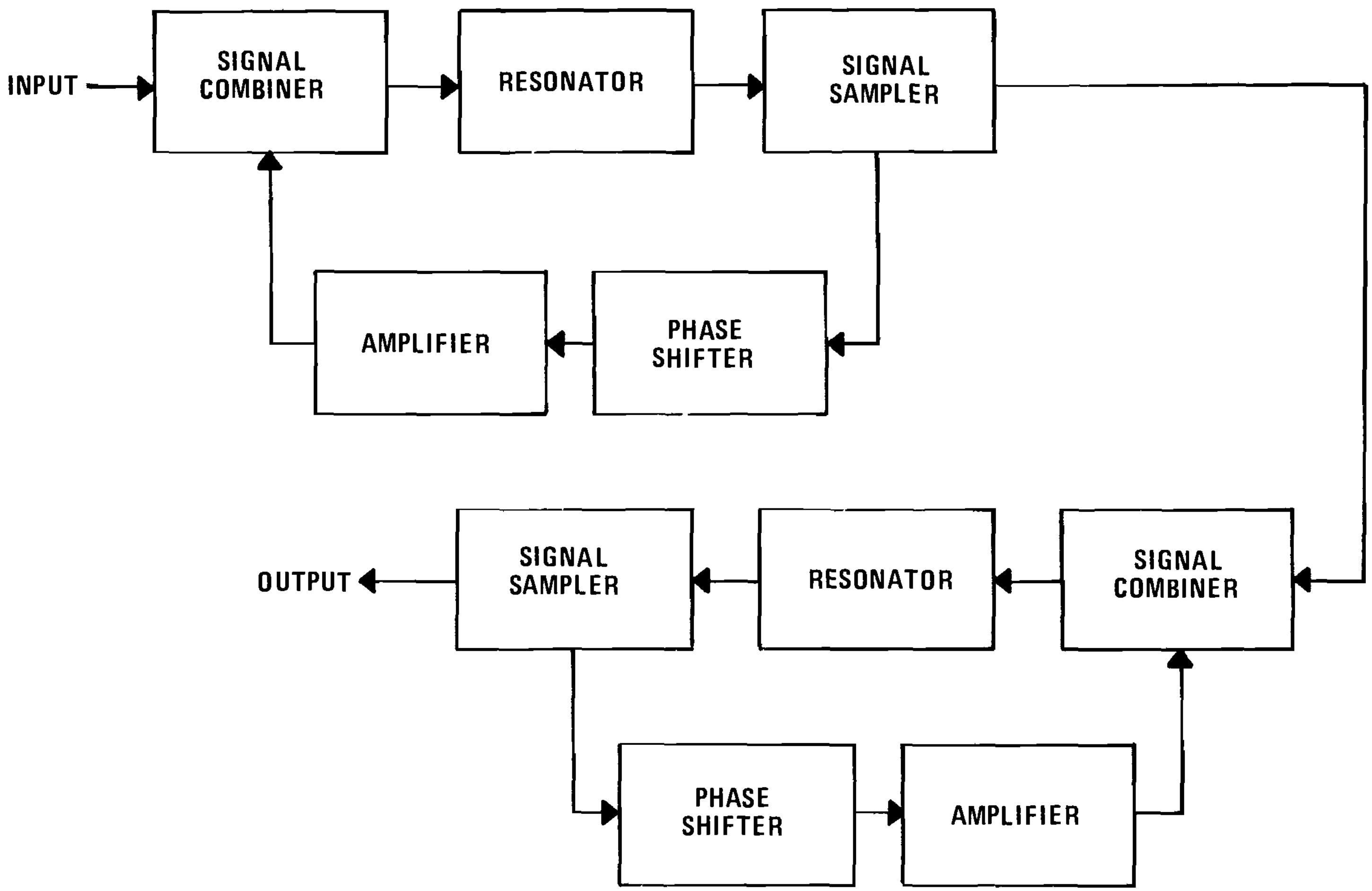


Figure 3. Block Diagram Showing the Cascade Operation of Resonators of Multiplied Q to Obtain a Double Tuned Response.

tuned for minimum bandwidth or they may be stagger tuned to produce a flat passband region with more rapid attenuation roll-off in the stop-band. The response characteristics of one, two, and three cascaded stages of Q multiplication are illustrated in Figure 4. Shown for comparison purposes are the response curves of a single cavity and a double cavity coaxial filter. Note that the 60 dB bandwidth of the two stage active filter is 9 MHz compared to the 12 MHz 60 dB bandwidth of the model 156C-2 double tuned cavity filter.

The effectiveness of the active filters for improving the interference rejection properties of a receiver is illustrated in Figure 5. The cross modulation rejection of an AN/APR-4 receiver with its internal preselector is compared with the rejection obtainable when various types of active and passive filters are added to the front end of the receiver. The curves show the significant improvement in the interference rejection which can be achieved with filters of multiplied Q. With only the internal preselector, the AN/APR-4 exhibits 30 dB rejection to an interfering signal spaced 2 MHz from the tuned frequency at 300 MHz. The single cavity filter increases this rejection to 52 dB and the model 156C-2 multicoupler provides an increase in the rejection to 70 dB. The three stage Q multiplier filter increases the rejection of the receiver to the interfering signal to 92 dB for an improvement over the basic receiver of 62 dB, and a 22 dB improvement over the double tuned passive multicoupler. The maximum cross modulation rejection available out of band to interfering signals was limited to approximately 90 dB with only the internal preselector of the receiver. The added active filters increased this ultimate rejection to greater than 110 dB.

B. Negative Resistance

Another well-known technique of Q multiplication is the use of active networks to produce a negative resistance which in turn supplies some of the energy losses in the resonator. The result is a lower net value of positive resistance and a consequent increase in Q.

A negative resistance device is characterized by

$$e = -iR \quad . \tag{1}$$

The minus sign indicates that the e-i characteristic curve has a negative slope. No physically realizable device can exhibit a negative e-i characteristic over its entire operating range. However, a number of common devices such as the tetrode vacuum tube and the tunnel diode exhibit limited regions of negative resistance. Properly biased within the negative resistance region, such devices can supply power to the input terminals. In other words, the power reflected from the device is greater than the power incident on the device.

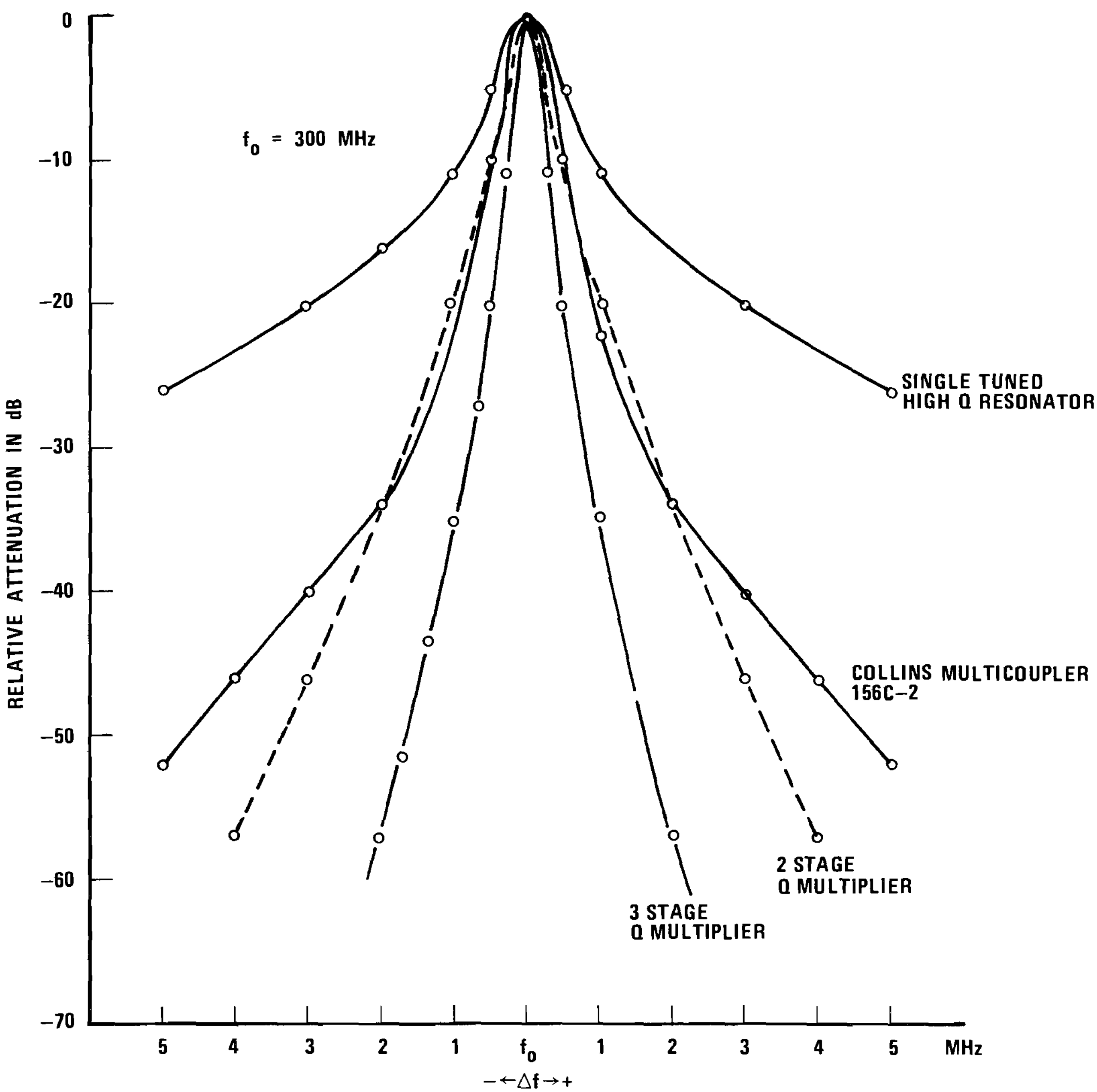


Figure 4. A Comparison of the Response Characteristics of Selected Active and Passive Preselector Types.

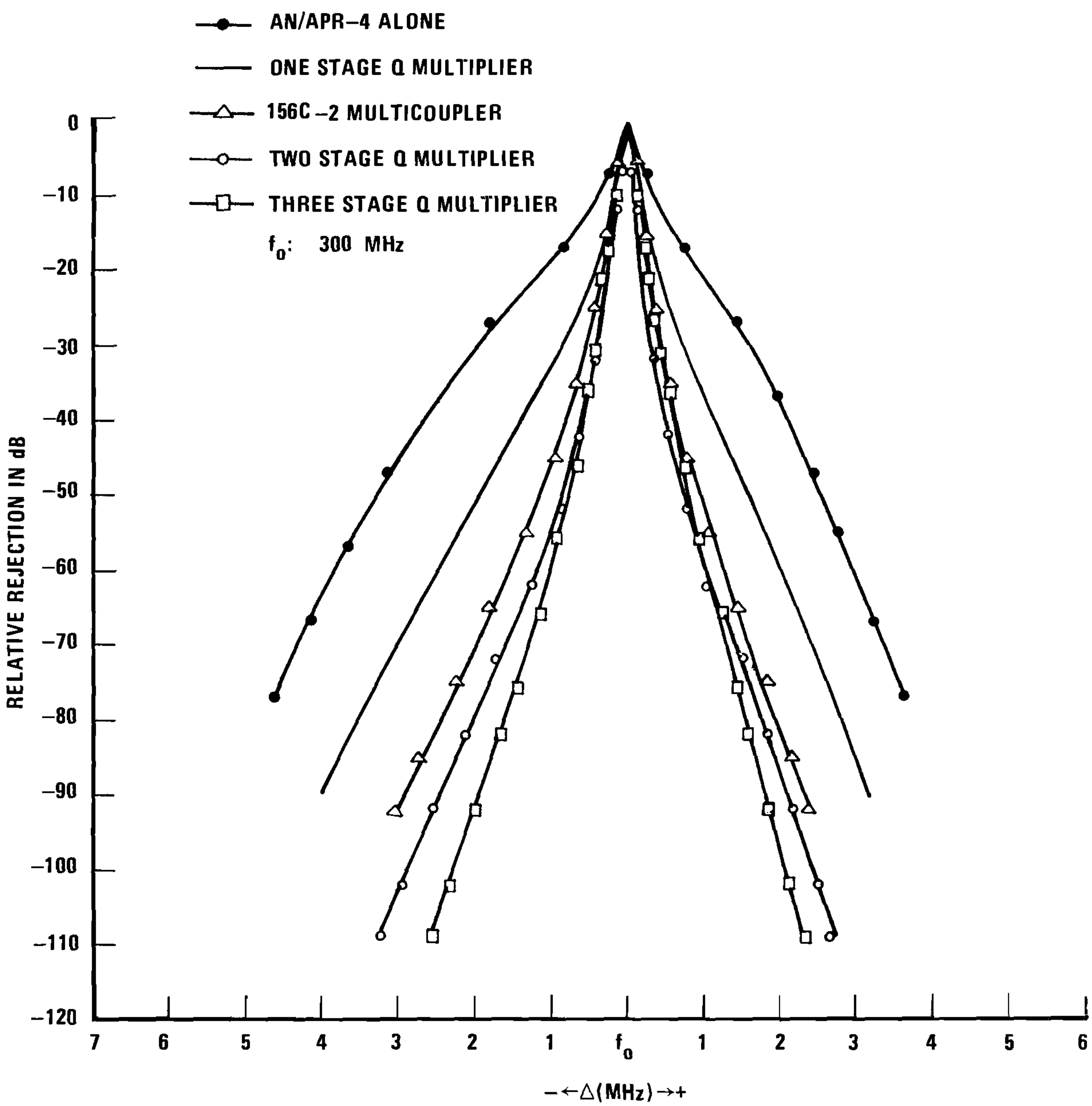


Figure 5. Cross Modulation Rejection Characteristics of a UHF Communications Receiver With Selected Active and Passive Filters Added.

If a transmission line of characteristic impedance Z_o is terminated in an impedance of Z_i, the ratio of the reflected voltage to the incident voltage, i.e., the reflection coefficient, is

$$\Gamma = \frac{Z_o - Z_i}{Z_o + Z_i} \quad . \tag{2}$$

Depending upon the ratio of the Z_o to Z_i, Γ may be positive or negative, but, for Z_i positive, its maximum value is one as shown by Figure 6. If Z_i is negative, however, Γ can have any value between plus and minus one to infinity. A reflection coefficient greater than one indicates that the terminating resistance is supplying power to the transmission line. Except for $Z_i = -Z_o$, the system is stable.

A true negative resistance is a two terminal device. Being a two terminal device, it may be connected in series with or in parallel to a positive resistance to reduce the net total resistance. To maintain stability, the net resistance must remain positive. Negative resistance devices exhibit either open circuit stability or short circuit stability [5]. An open-circuit stable device is controlled by the current passing through the terminals of the device and, as such, is appropriate for a series mode of operation. A short circuit stable device is controlled by the voltage applied across the terminals of the device and thus is appropriate for a parallel mode of operation. Parallel mode devices are of particular interest to UHF Q multiplication applications. They are simple to implement in that they may be used as terminating elements of transmission line sections. The terminated transmission lines may be connected as shunt elements as shown in Figure 7. They may also be coupled into cavity walls or connected across resonant lengths of line to multiply the Q of resonators directly.

The use of tunnel diodes as negative resistance terminating elements is well known. For example, a tunnel diode may be employed as the terminating element of a low-pass filter [6] to produce a low noise, low-pass amplifier. The tunnel diode finds its primary usefulness in low signal level applications. The limited signal handling capabilities of tunnel diodes restrict their application as Q multiplication devices.

A common collector transistor circuit has been suggested as exhibiting a negative resistance characteristic [7]. On the basis of its demonstrated performance at low frequencies, further investigation of its characteristics in the 225 to 400 MHz range was performed. Note that it is not correct to refer to the common collector amplifier as an emitter follower at these frequencies because the emitter voltage does not really "follow" the base voltage as it does at low frequencies. As a negative resistance device, the common collector circuit offers a significant advantage over the tunnel diode in power handling capability.

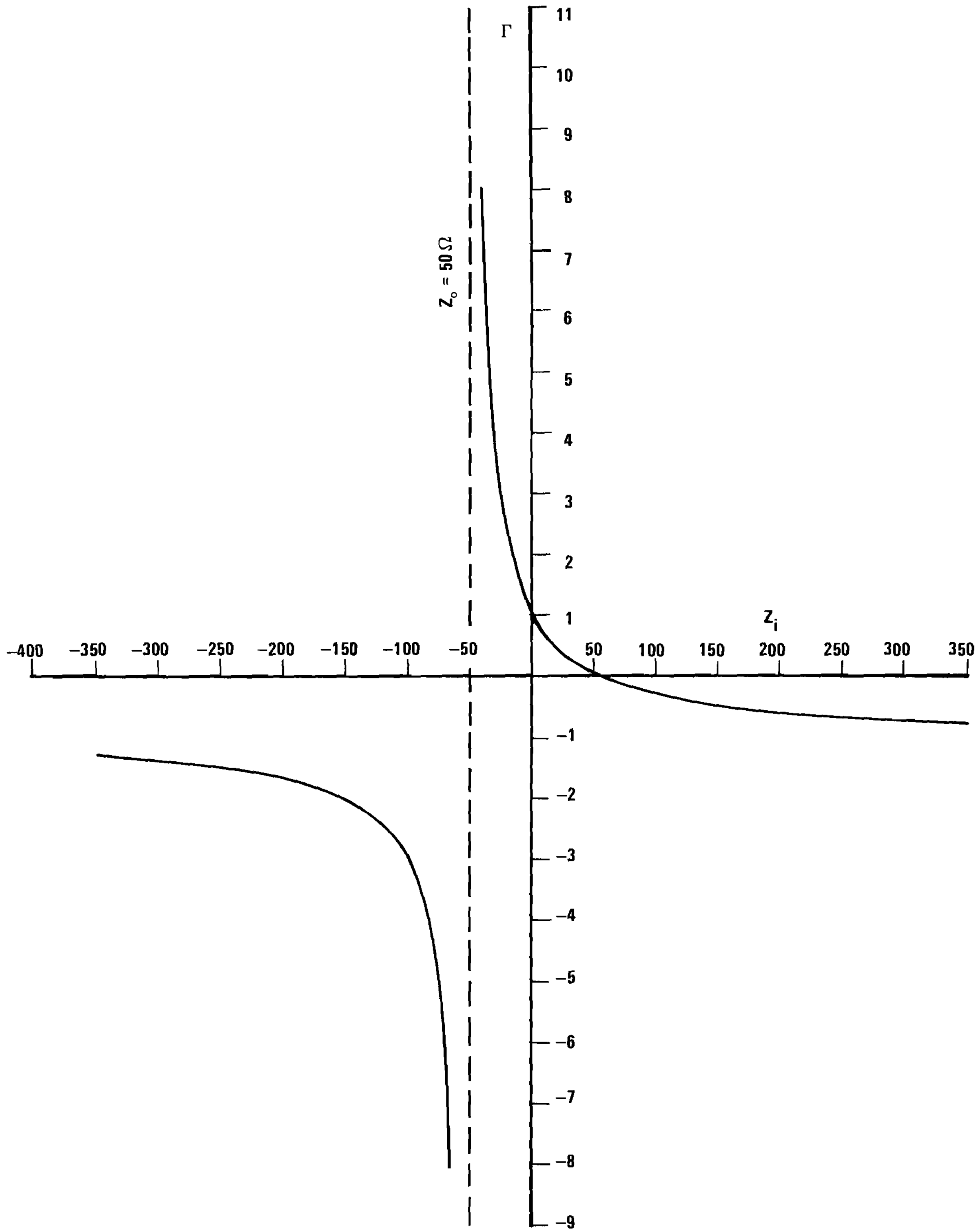


Figure 6. Reflection Coefficient as a Function of Both Positive and Negative Terminating Impedances.

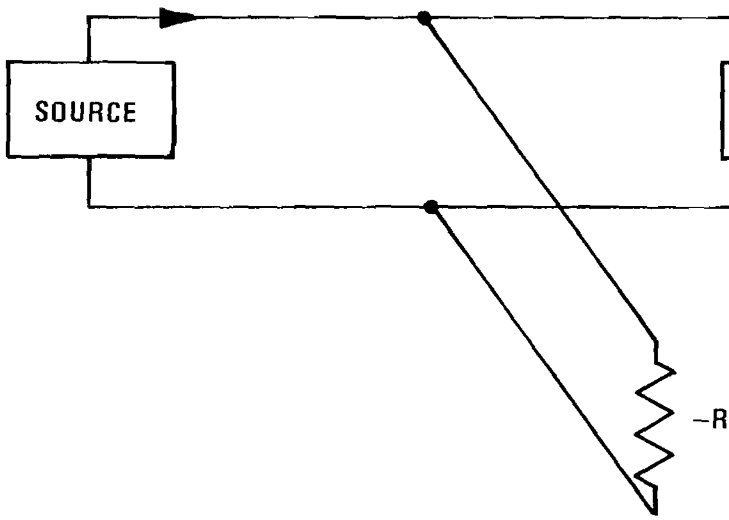


Figure 7. Illustration of the Application of Negative Resistance as a Shunt Element Across a Transmission Line.

Consider the grounded collector amplifier shown in Figure 8. Measurements of the scattering parameters of this amplifier reveal its characteristics at 300 MHz to be

$$S_{11} = 0.8 \angle -60^{\circ} \; ; \tag{3}$$

$$S_{22} = 0.83 \angle +95^{\circ} \; ; \tag{4}$$

$$S_{21} = 1.66 \angle -75^{\circ} \; ; \tag{5}$$

$$S_{12} = 0.52 \angle +9^{\circ} \; . \tag{6}$$

The terminal characteristics of the amplifier with an arbitrary load are most readily obtained through the use of signal flow graph analysis techniques. The flow graph of a four terminal network driven by a source, b_s, with a reflection coefficient, Γ_s, and terminated with a load having a reflection coefficient, Γ_L, is shown in Figure 9. If Γ_s is assumed to be zero (i.e., the impedance of b_s is equal to the measurement reference impedance which in this case is 50 ohms), the power flow $b_s b_1$ is equal to the power flow $a_1 b_1$.

The effective input reflection coefficient of the amplifier for the load not matched is [8]

$$T_{a_1 b_1} = \frac{\Sigma T_k \Delta_k}{\Delta} \quad , \tag{7}$$

where

T_k = the gain of the kth forward path,

Δ_k = the value of Δ not touching the kth forward path, and

Δ = 1 - (sum of all individual loop gains) + (sum of the loop gain products of all possible combinations of two non-touching loops) - (sum of the loop gain products of all possible combinations of three non-touching loops) + ...

Thus, for $\Gamma_s = 0$, the effective input reflection coefficient, S_{11}', is

$$S_{11}' = T_{a_1 b_1} = S_{11} + \frac{S_{12}\, S_{21}\, \Gamma_L}{1 - S_{22}\, \Gamma_L} \quad . \tag{8}$$

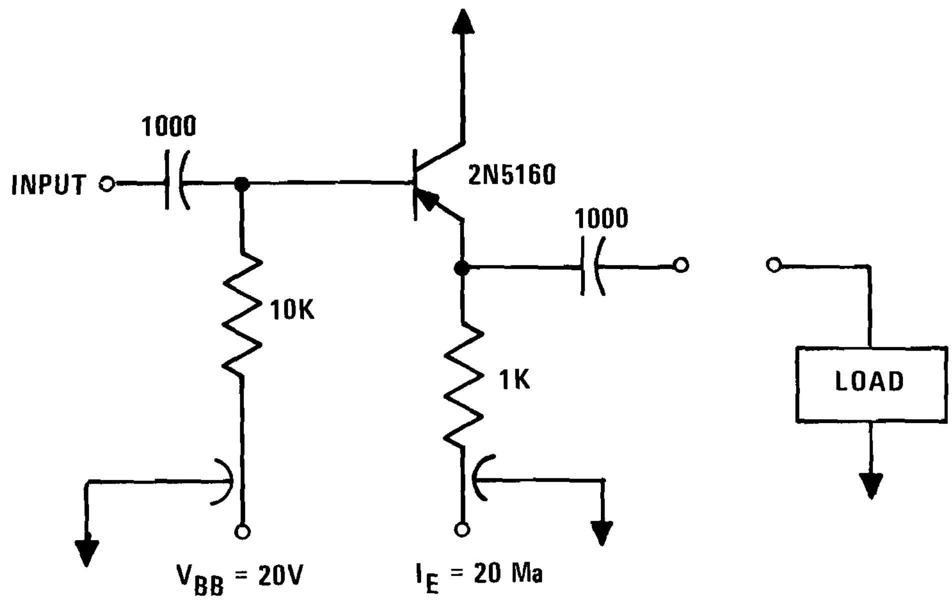


Figure 8. Schematic of a Grounded Collector Amplifier.

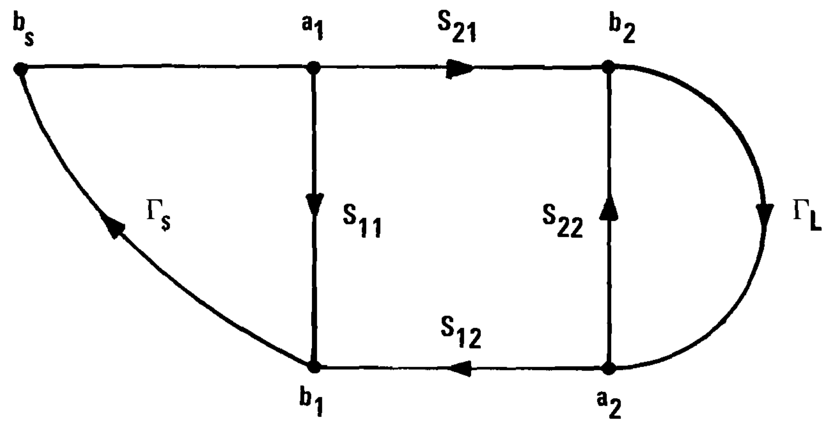


Figure 9. Signal Flow Graph for a Simple Amplifier.

For the transistor with a realizable load

$$|S_{12}S_{22}\Gamma_L| \geq 0 \quad , \tag{9}$$

and

$$|S_{22}| \text{ and } |\Gamma_L| \text{ are } \leq 1 \quad . \tag{10}$$

For stability, it is necessary that

$$S_{22}\,\Gamma_L < 1 \quad . \tag{11}$$

It may be seen from (8), that the circuit will oscillate only if both S_{22} and Γ_L are equal to one. In particular, if S_{22} is less than one at all frequencies, stability is assured.

Substitution of the measured parameters (3)-(6) into (8) yields

$$S'_{11} = 0.8\,\angle{-60^\circ} + \frac{(0.86\,\angle{-66^\circ})\,\Gamma_L}{1 - (0.83\,\angle{+95^\circ})\,\Gamma_L} \quad . \tag{12}$$

Assume a pure capacitive load having a Γ_L of -j1. Then

$$S'_{11} = -\,2.62 - j3.99 = 4.8\,\angle{-125^\circ} \quad . \tag{13}$$

From (13), it is evident the circuit exhibits a reflection coefficient greater than one and therefore it behaves as a negative resistance in that it supplies more power to the input terminals than was originally supplied by the source.

Figure 10 illustrates the use of a common collector negative resistance amplifier to enhance the Q of a coaxial cavity. Since the source impedance consists of the parallel combination of the cavity output impedance and the other terminal impedance, the source reflection coefficient, Γ_s, is not zero except possibly at specific frequencies. If $\Gamma_s \neq 0$, the conditions for stability are not as well defined as implied by equation 11. The amplifier can be made stable regardless of the tuning point of the cavity by adjusting the emitter load impedance as shown in Figure 11 and by appropriately choosing the cavity output coupling configuration. A high impedance probe was experimentally determined to be the optimum cavity coupling for best Q commensurate

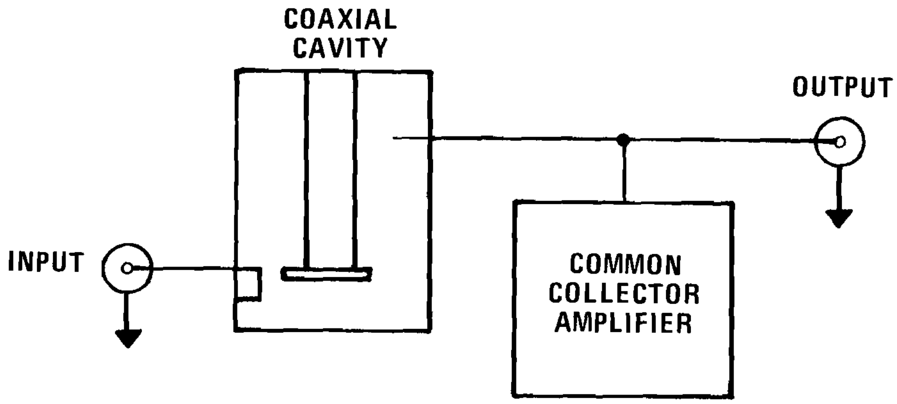


Figure 10. Application of the Negative Resistance Amplifier to the Multiplication of the Q of a Coaxial Cavity.

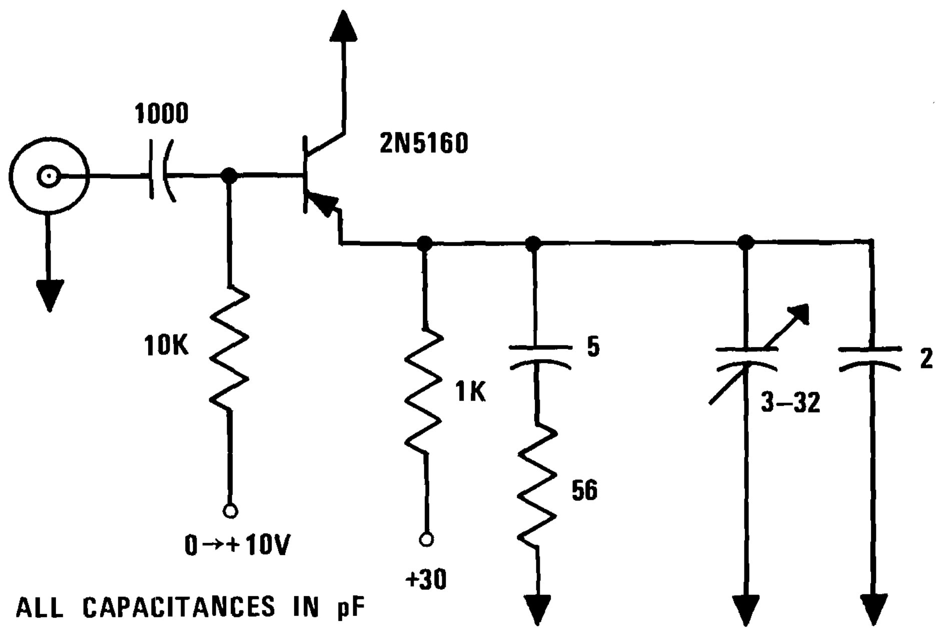


Figure 11. Circuit Diagram of a Stable Negative Resistance Amplifier.

with stable operation over a wide frequency range. Since the transistor characteristics are not constant with frequency, the emitter load capacitor must be trimmed slightly at different frequencies to obtain maximum multiplication.

The effectiveness of the common collector amplifier as a Q multiplier is illustrated by Figure 12. The Q of a cavity with and without Q multiplication is shown as a function of frequency. The cavity terminal impedance as well as the transistor parameters vary with frequency and, consequently, the relative multiplication is frequency dependent.

The common collector amplifier in Figure 11 can typically handle signal levels up to -5 dBm. By using the emitter as the input port with the base as the "output" port, an increase in power handling capability may be expected. This "Inverted Common Collector" configuration has been investigated independently by Adams and Ho [9]. They have shown that this configuration also exhibits negative resistance properties in the UHF frequency region.

C. Q Multiplier Filter

The breadboard model of a tunable bandpass filter that is shown in Figure 13 was constructed to demonstrate the feasibility of active filter techniques in the 225 to 400 MHz region. A combination of negative resistance Q multiplication and lightly loaded coaxial cavities was employed to produce a narrow bandpass response with high skirt selectivity. The negative resistance technique of Q multiplication was chosen because it offered the capability of operating over the entire frequency range. At the time of construction of the filter, a phase shifter capable of operating from 225 to 400 MHz was not available and, therefore, the positive feedback technique could not be implemented over the total range.

The curves of Figure 4 indicate that cascaded resonators are necessary to obtain high skirt selectivity. Single stage Q multiplication achieves reduction of the 3 dB bandwidth and, by virtue of the available gain, produces a relative improvement in stop band attenuation, but it does not increase the slope of the skirt attenuation characteristic. Consequently, in addition to the Q multiplication stage, two passive stages are included to provide high skirt selectivity. Additional amplifiers are provided to establish the noise figure of the system and/or to provide isolation between the tuned elements. Two selectable attenuators of the design shown in Figure 14 are included to limit the gain in certain frequency regions and to provide impedance stabilization for the negative resistance amplifier.

The active and passive elements are combined in the system shown in the block diagram of Figure 15 to produce a bandpass filter response having a 3 dB bandwidth of 200 ±50 kHz, a 6 dB-to-60 dB selectivity ratio of 0.07 and a skirt rolloff attenuation slope of 18 dB/octave. A minimum of 30 dB gain is available at any tuned frequency between 225 and 400 MHz. The noise figure of the filter is approximately 20 dB.

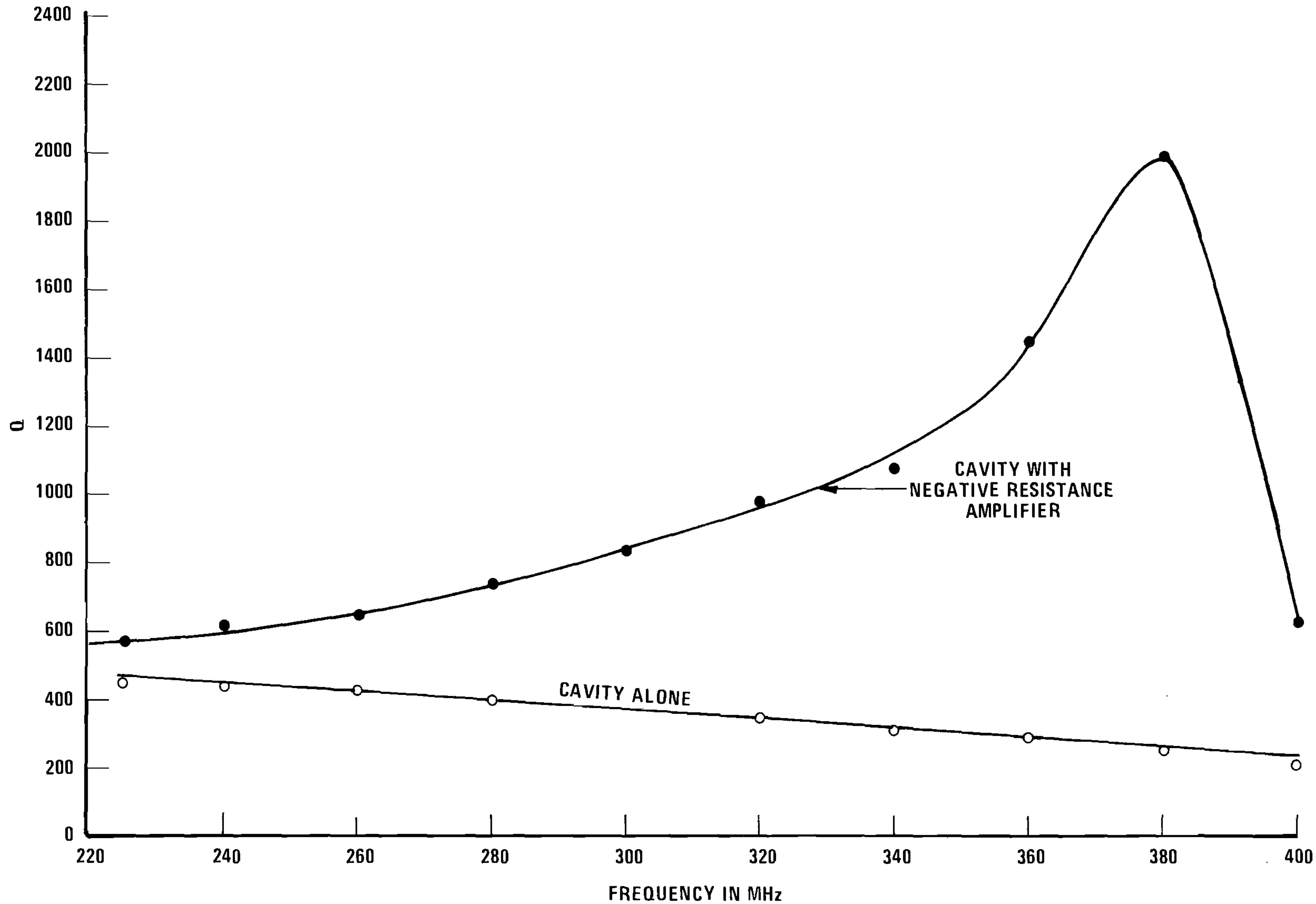


Figure 12. Cavity Effective Q as a Function of Frequency with Negative Resistance Q Multiplication Applied.

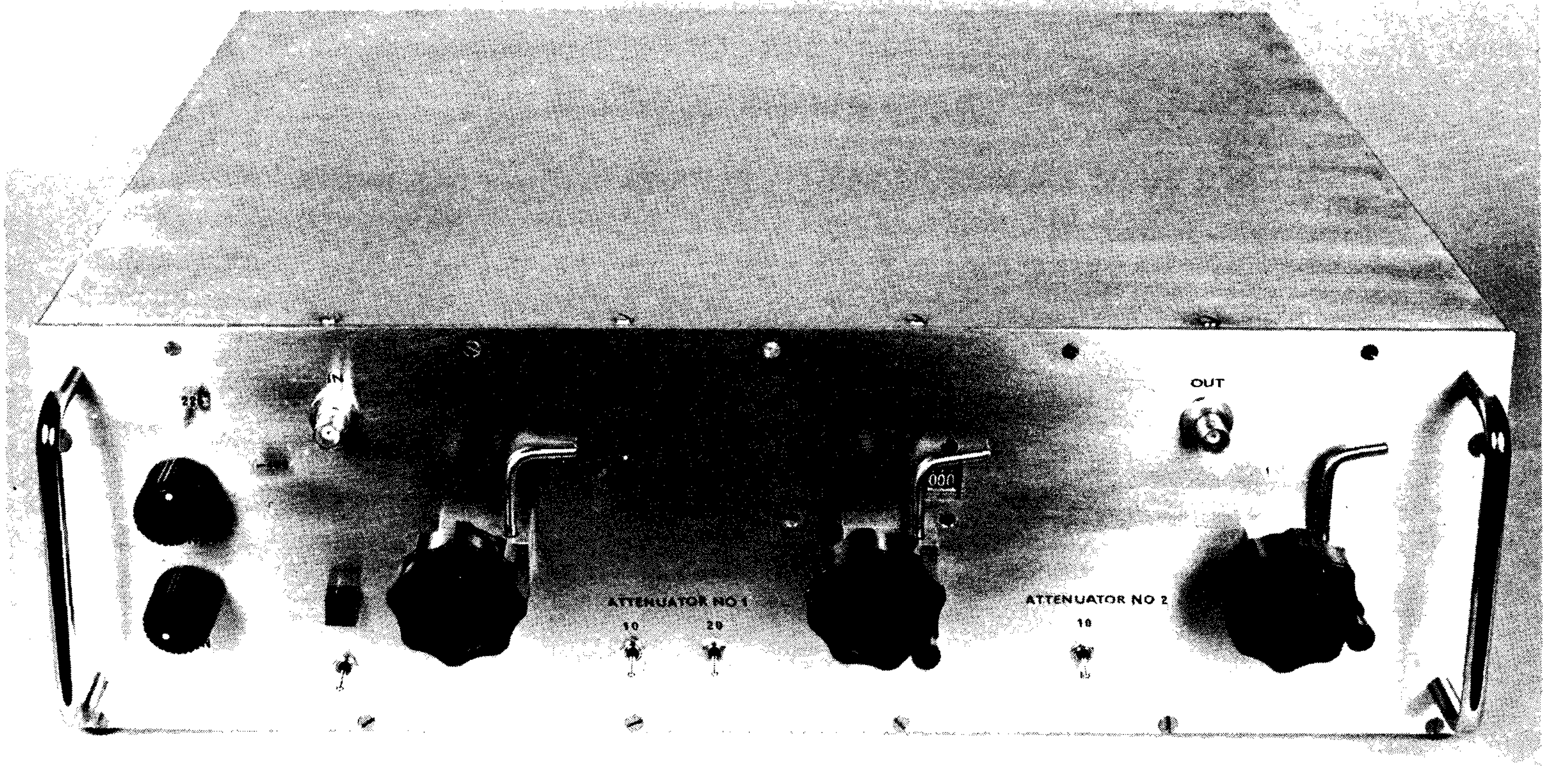


Figure 13. Photograph of the Cascaded Q Multiplier Filter.

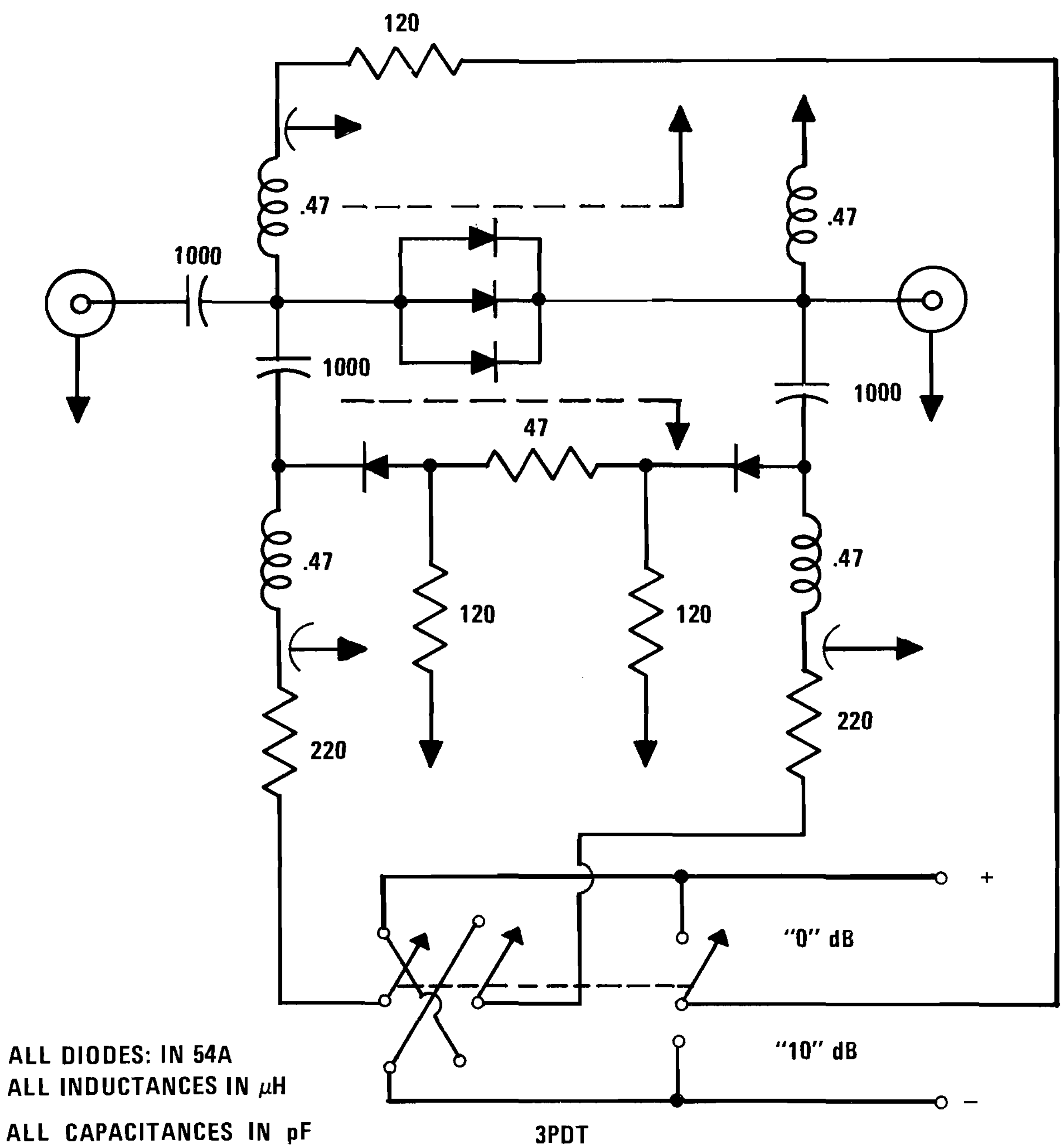


Figure 14. Schematic Diagram of Electronically Switchable Broadband Attenuator.

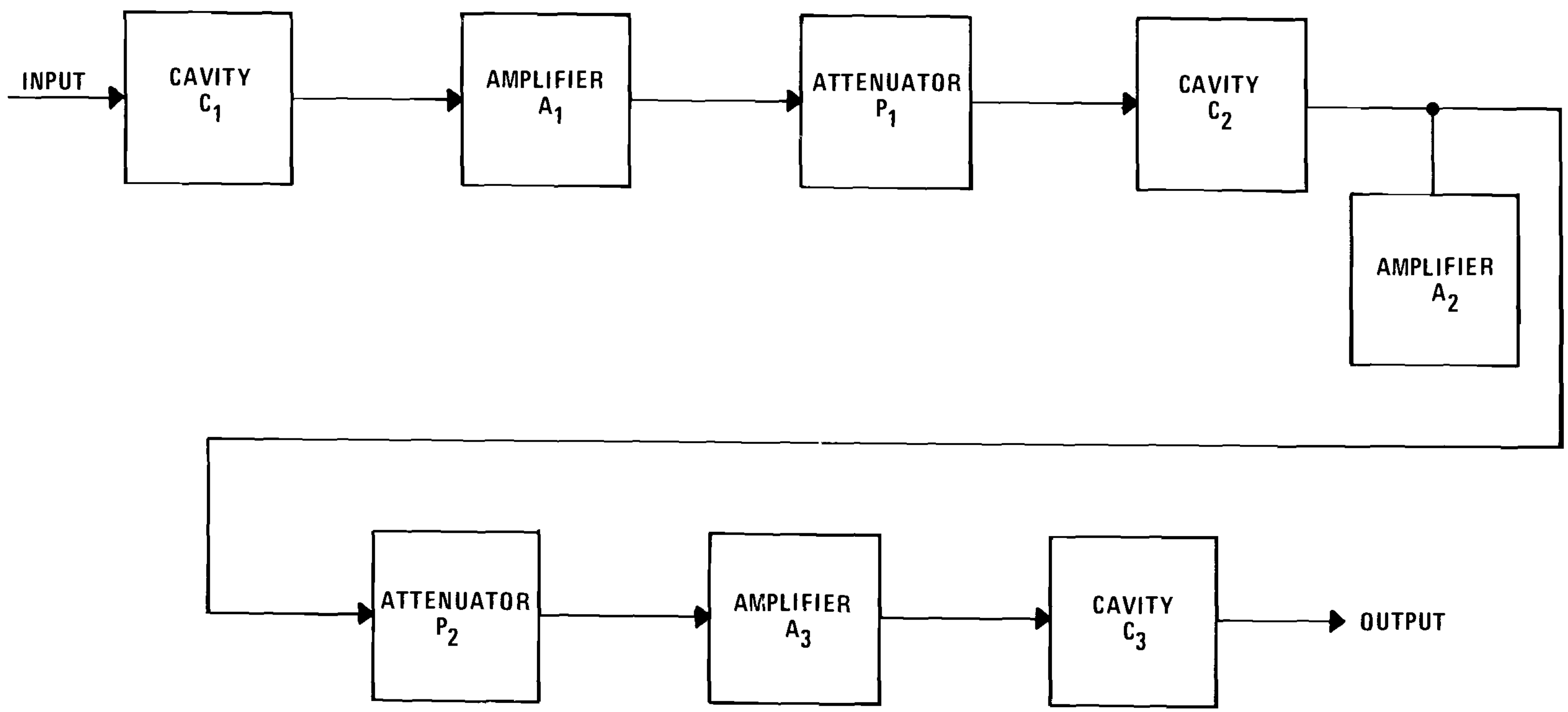


Figure 15. Block Diagram of the Cascaded Q Multiplier Filter.

<u>System Description</u>: The input signals from the antenna are first filtered by cavity, C_1. The coupling loops in C_1 are adjusted to provide a moderate loaded Q of approximately 500 with an insertion loss of typically 5 dB across the frequency range.

A broadband amplifier, A_1, provides approximately 28 dB gain to effectively establish the noise figure of the filter. The tangential sensitivity of A_1 is about -105 dBm and the 1 dB compression level is -5 dBm for a dynamic range of approximately 100 dB. The input VSWR is less than 1.5:1 and the output VSWR is less than 2.0:1 over the frequency range.

Attenuator P_1 can be set at 0, 10, and 20 dB with front panel switches. P_1 provides a coarse gain adjustment for the filter and adds additional impedance stabilization to the second cavity, C_2.

The input coupling to C_2 is a small inductive loop which is adjusted to provide optimum Q commensurate with moderate insertion loss at the low end of the frequency range. Output coupling is provided with an E-field probe whose length and position are adjusted to provide the proper impedance for stable operation of amplifier A_2. Amplifier A_2 is a common collector configuration which operates in parallel to the output port of C_2 and provides effective multiplication of the Q of the cavity.

P_2 is an attenuator whose value is selectable at either 0 or 10 dB. The 10 dB position is necessary to assure stability of A_2 in the frequency region of 350-400 MHz.

To provide a high degree of isolation between A_2 and cavity C_3 over the entire band, amplifier A_3 is provided. A_3 is a broadband amplifier which provides the required isolation as well as supplying approximately 20 dB gain.

Since the net gain from the input of C_1 to the output of A_3 is high across the band, high insertion loss can be tolerated in C_3. Therefore, small coupling loops oriented to realize the maximum possible Q are used in C_3.

The tangential sensitivity of the filter across the 225 to 400 MHz range is shown in Figure 16. A 5 to 10 dB improvement in the sensitivity could be realized by reducing the insertion loss of cavity C_1 and improving the noise figure of amplifier A_1. The effective filter Q as a function of frequency is shown in Figure 17. Note that higher Q's are evident between 300 and 400 MHz which is to be expected because of the higher multiplication provided by the negative resistance amplifier in this region.

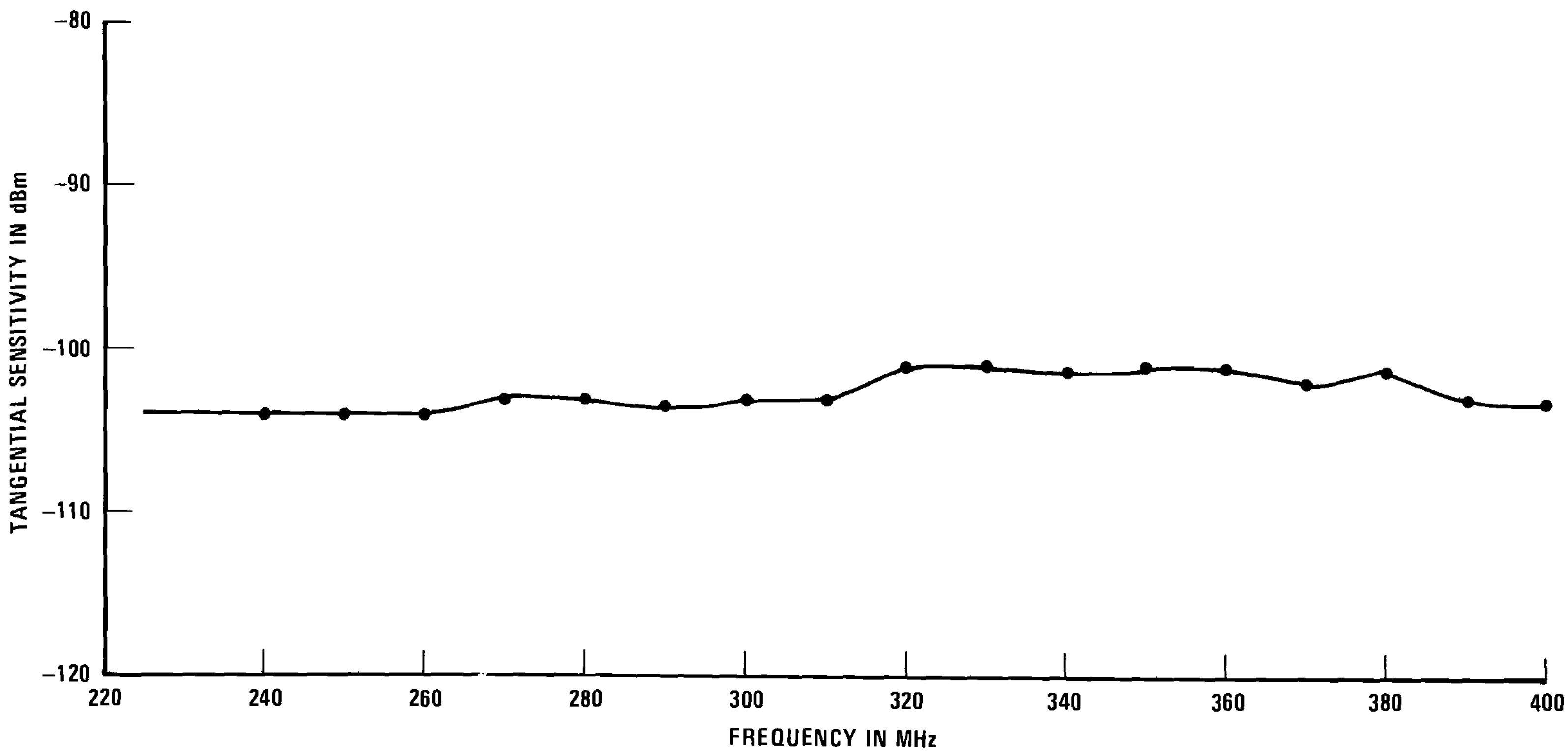


Figure 16. Tangential Sensitivity of the Cascaded Q Multiplier Filter.

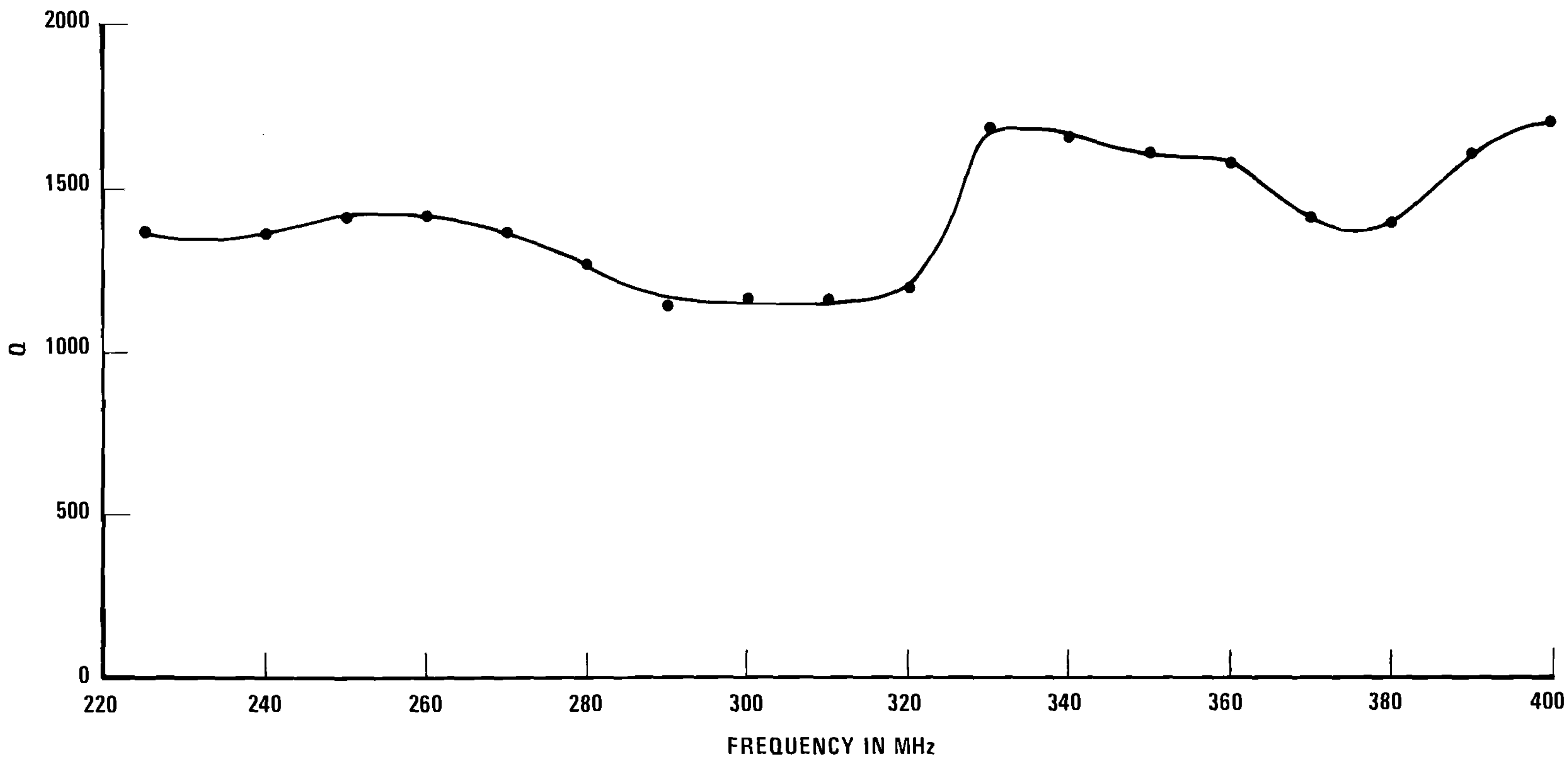


Figure 17. Effective Q of the Q Multiplier Filter as a Function of Frequency.

SECTION III

AM CANCELLATION FILTER

A. Introduction

Closely spaced unwanted signals can present a serious problem in communication receivers. If the level of the interfering signal is small, the interference may be manifested as an annoying beat frequency between the two signals which seriously degrades the intelligibility of the desired signal. In the case of a high level interfering signal, more serious problems often arise due to intermodulation product generation or complete saturation of the receiver RF amplifiers and mixers.

Conventional methods for overcoming these problems include the use of high Q bandpass filters to pass only the desired signal or the use of band reject filters to suppress the unwanted signal. Such methods are effective for signals adequately spaced in frequency. However, if the interfering signal is very close in frequency to the desired signal, then a conventional bandpass filter will not offer sufficient rejection to the unwanted signal while the band reject filter will excessively attenuate the desired signal.

A method of circumventing the limitations of conventional filters is to suppress the interference with a signal whose amplitude and phase are identical to that of the interfering signal. Subtracting this replica of the interfering signal from the sum of the desired and interfering signals results in cancellation of the interference leaving only the desired signal. When it is possible to obtain a sample of the interference directly as from a cosited transmitter through the use of a coaxial cable connection, then the phase and amplitude of the sample can be properly adjusted to produce cancellation of the interference. If a sample of the interfering signal is not available, it is then necessary to resort to the more complex approach of synthesizing a replica of the interfering signal.

A convenient method for synthesizing this cancellation signal is to generate an auxiliary signal which can be phase-locked to the interfering signal. This phase-lock technique is employed in the UHF AM Cancellation Filter to suppress both CW and amplitude modulated signals in the 200 to 400 MHz frequency range.

The UHF AM Cancellation Filter shown in Figure 18 is essentially a dual channel, double conversion, superheterodyne receiver with a 200-400 MHz cancellation oscillator and the necessary phase and amplitude control loops added. Basically, the system employs a common local oscillator to heterodyne separately the interfering signal and the cancellation signal to a convenient IF frequency. The two IF signals are generated in separate mixers and delivered to their respective IF amplifiers.

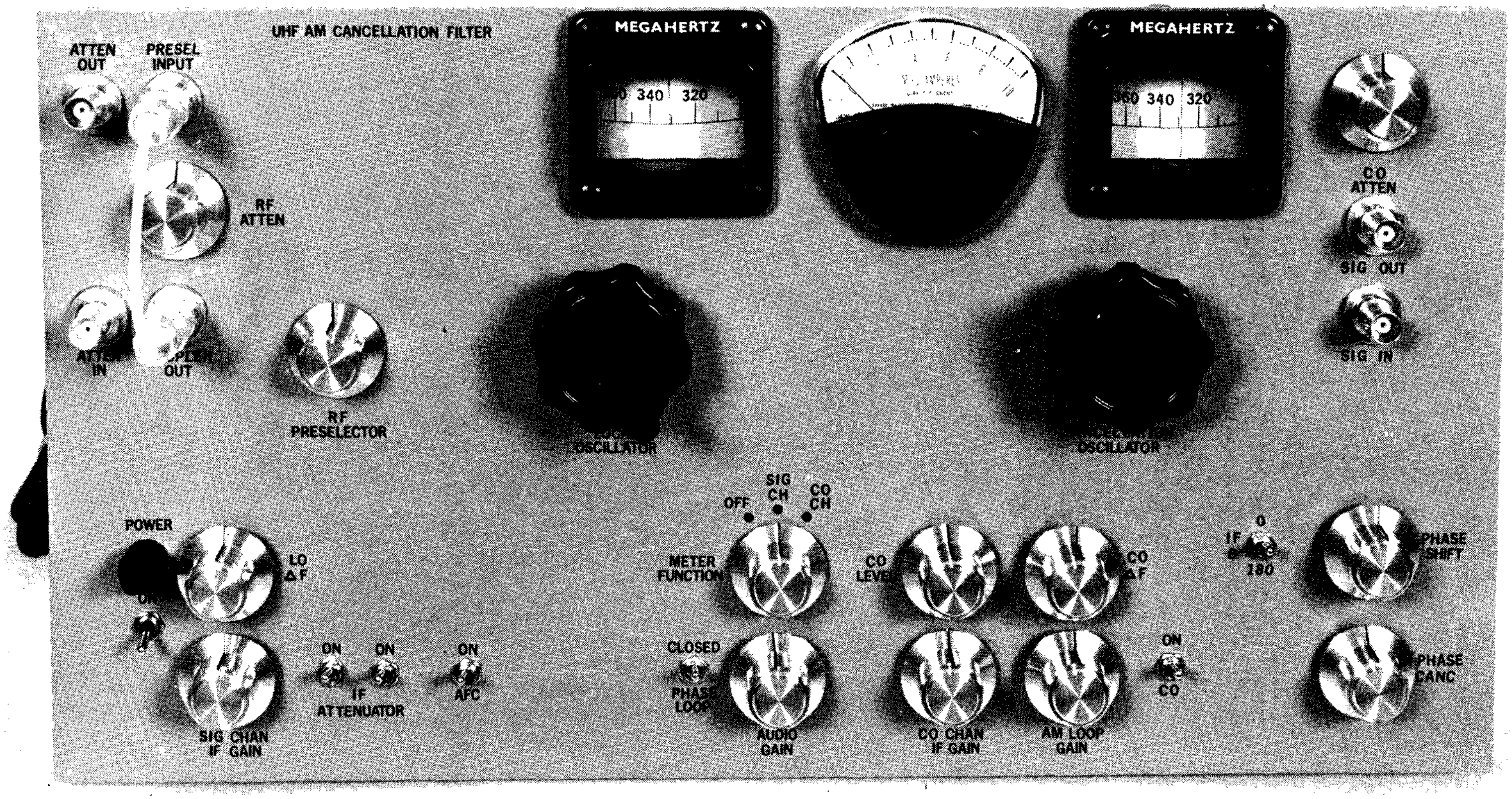


Figure 18. Front Panel View of the UHF AM Cancellation Filter.

The IF signals are then compared in a phase detector and an amplitude detector; the resulting error signals are used to control the frequency and amplitude, respectively, of the cancellation signal. The cancellation oscillator signal, being properly adjusted in frequency, phase and amplitude, is summed with the interference in a hybrid combiner to suppress the undesired signal.

B. Circuit Description

A block diagram of the complete UHF AM Cancellation Filter is given in Figure 19. Referring to the block diagram, the input signal which consists of both the desired and undesired signals is connected to the hybrid combiner and then the output of the hybrid is connected to the 10 dB directional coupler. A portion of the input signal is fed to the RF attenuator and RF preselector. The continuously variable RF attenuator is provided to prevent high level signals from overdriving the preselector. The signal is amplified by the transistor RF preselector and delivered to the first mixer in the signal channel. Concurrently, the low level output from the cancellation oscillator is delivered to the first mixer in the cancellation channel. Two isolated outputs from the first local oscillator are fed to each of the first mixers where the input signal and the cancellation signal are heterodyned to 31 MHz and delivered to their respective first IF amplifiers. The 31 MHz signals from each of the first IF amplifiers are delivered to the second mixers where they are heterodyned to a 7 MHz second IF by the crystal controlled second local oscillator.

The signal channel, 7 MHz IF output is delivered simultaneously to the phase detector, amplitude detector, envelope detector, IF limiter and to a rear-panel output connector. The envelope-detected signal provides the drive for the audio amplifier and the meter amplifier. The output of the IF limiter is coupled to a modified Foster-Sealey discriminator which provides the AFC voltage for the first local oscillator.

The cancellation channel, 7 MHz IF output is delivered to the variable phase shifter. A portion of this signal is also delivered to a rear panel output connector. The 7 MHz signal is also envelope detected to provide a level indicator. The voltage controlled variable phase shifter provides greater than 360 degrees phase adjustment and is used to set the loop phase to the value required for a stable operating condition. The output of the variable phase shifter is split into two quadrature signals for application to the phase and amplitude detectors.

The phase detector output is filtered by the lag network and applied to a varactor diode in the cancellation oscillator circuit. The phase error voltage adjusts the frequency of the cancellation oscillator to maintain phase-lock.

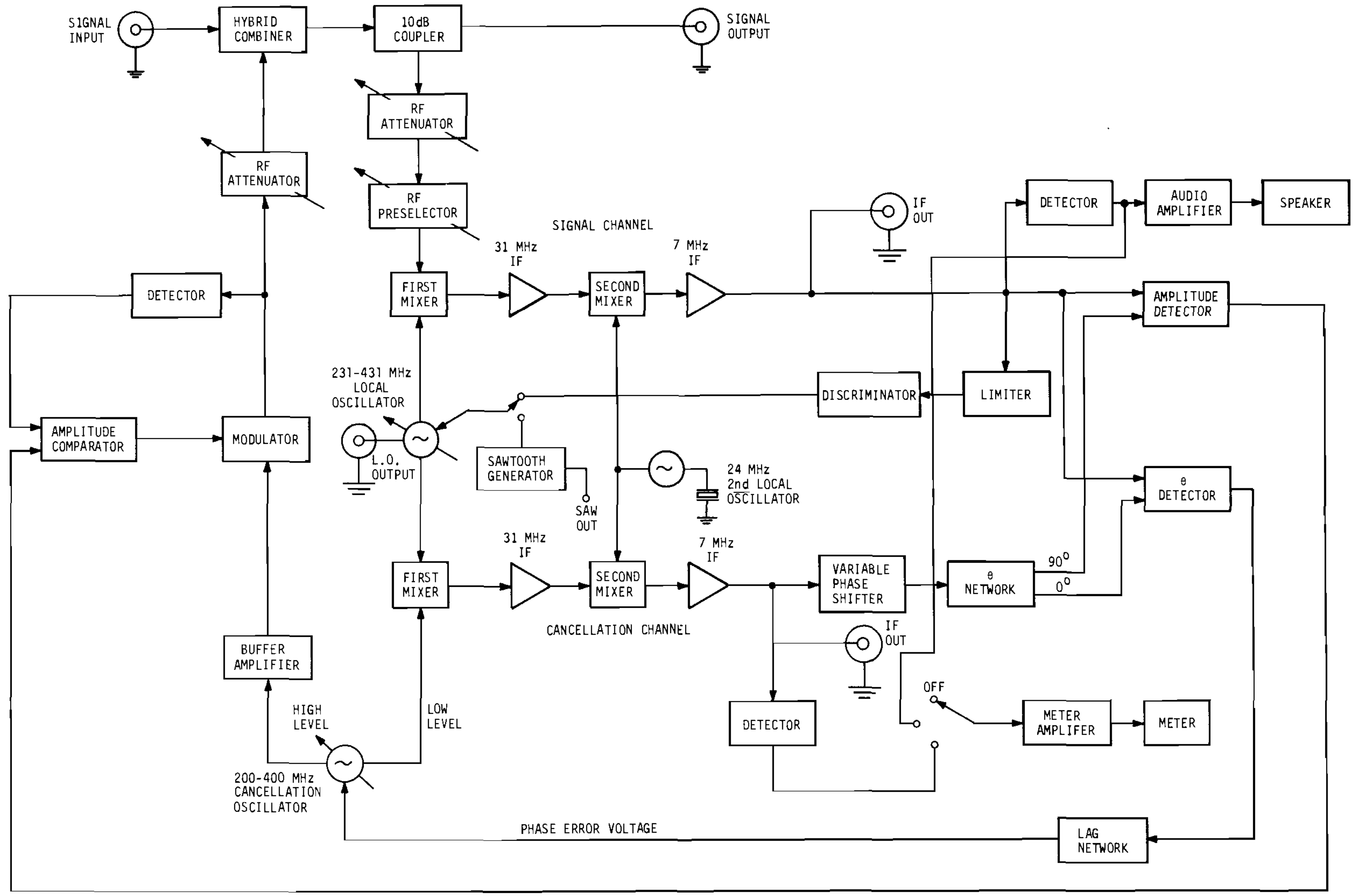


Figure 19. Block Diagram of the UHF AM Cancellation Filter.

When the cancellation oscillator is phase-locked to the interference, the error voltage produced by the amplitude detector represents the original modulation contained in the interfering signal. This signal is fed to the amplitude comparator whose output drives the PIN modulator. The high level output from the cancellation oscillator is delivered to the modulator through the buffer amplifier. The RF input impedance of the modulator varies as a function of the modulating signal which can introduce incidental FM into the cancellation signal because of the variable loading of the cancellation oscillator. The broadband buffer amplifier provides sufficient isolation between the oscillator and the modulator to prevent this incidential FM of the cancellation signal. A portion of the modulated cancellation signal is detected and compared with the error signal derived from the amplitude detector. This feedback modulator insures that the modulation of the cancellation signal will be a faithful reproduction of the error signal as obtained from the amplitude detector.

The cancellation signal, now properly adjusted in frequency, phase and amplitude, is summed with the desired and undesired signal to produce cancellation of the undesired interfering signal. It should be noted that the actual RF input to the cancellation filter, the -10 dB port of the directional coupler, follows the hybrid combiner. With this configuration, the cancellation filter operates on the residual of the interfering signal as a closed loop system. This technique also prevents the RF section of the cancellation filter from being overloaded by the large interfering signal once phase-lock is established and suppression is obtained.

Since the frequencies of the cancellation signal and the interfering signal are the same in a phase-locked condition, it is important that the center frequency response of the two IF channels be closely matched. A low frequency sawtooth generator is incorporated into the system to aid in the alignment of the two IF channels. For alignment, the sawtooth sweep voltage is fed to the varactor in the first local oscillator and an external RF signal is delivered to each of the first mixers. Alignment is then accomplished by matching the swept responses of the two IF channels.

C. Cancellation Capabilities

The ability of the UHF AM Cancellation Filter to suppress AM and CW interference is demonstrated by the following examples. The spectrum presentation of Figure 20A shows a large interfering signal that is 50 per cent amplitude modulated at 1 kHz, which almost completely obscures the low level desired signal which is located 100 kHz from the interference. When the two signals are processed by the cancellation filter, the relative levels are as shown in the photograph of Figure 20B. The interfering signal has been suppressed almost 40 dB and the desired signal is no longer obscured by the interfering signal. Although the interfering signal is still present, the level has been sufficiently

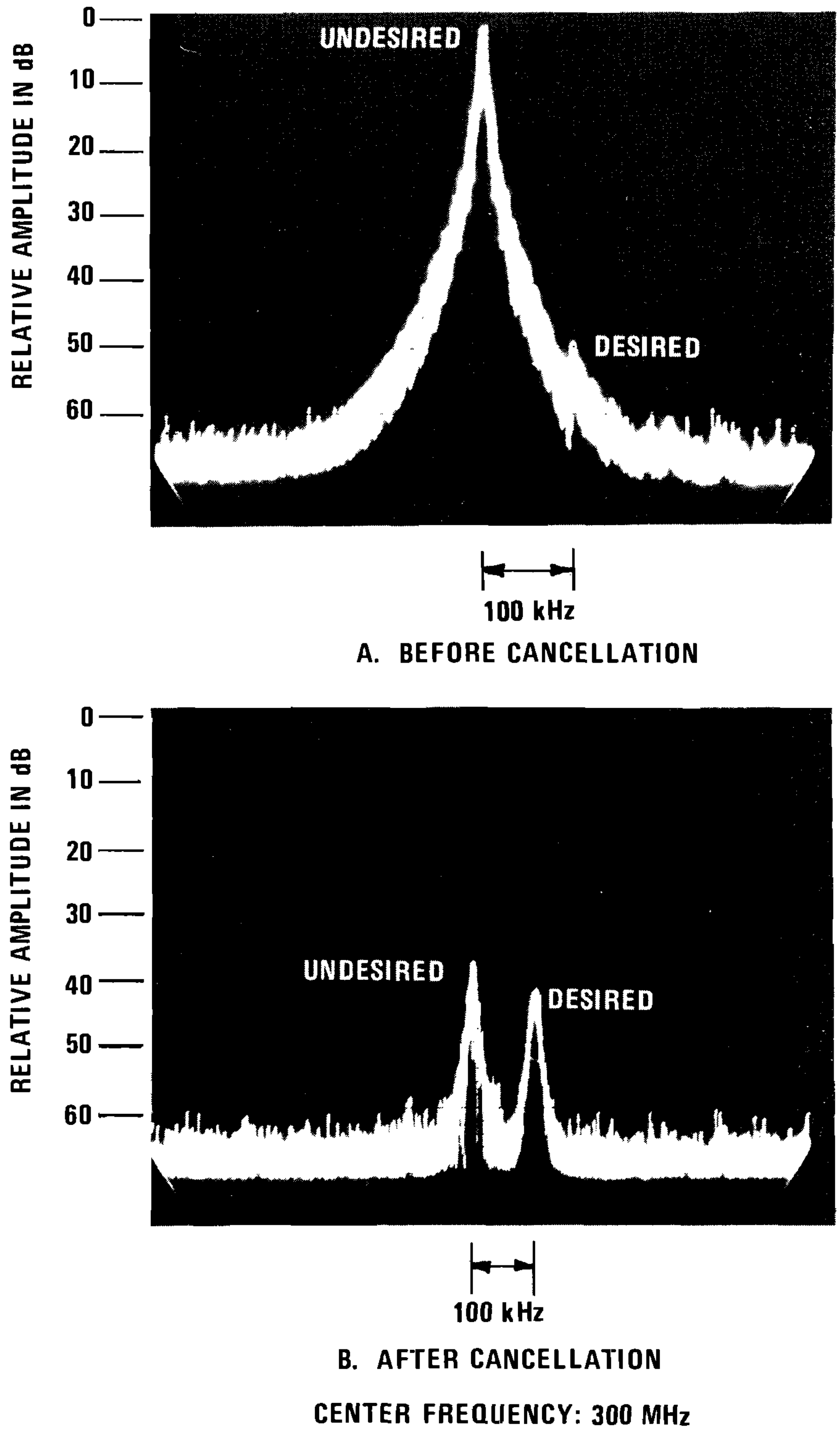


Figure 20. Spectrum Analyzer Displays Which Show Cancellation of an AM Signal With the AM Cancellation Filter.

reduced such that it would probably not overload a receiver front end. In this particular case, the selectivity of the receiver would, in general, provide the necessary additional rejection of the undesired signal.

In the spectrum display shown in the photograph of Figure 21A, the low level desired signal is almost concealed by a large CW interfering signal which is spaced in frequency only 30 kHz from the desired signal. In the spectrum presentation of Figure 21B, the interfering signal has been reduced approximately 50 dB and the desired signal is no longer hidden by the interfering signal.

As shown by these spectrum displays, the UHF AM Cancellation Filter is capable of providing 35 to 40 dB suppression of an interfering AM signal and 40 to 50 dB for CW interference. This difference can be attributed to the fact that for CW interference, the cancellation signal is required to track only the slow frequency and phase variations that occur between the interfering and cancellation signals. For an AM interfering signal, however, the cancellation signal must not only track the frequency variations but must also accurately reproduce the AM signal and track the incidental phase modulation that is generally present in amplitude modulated signals.

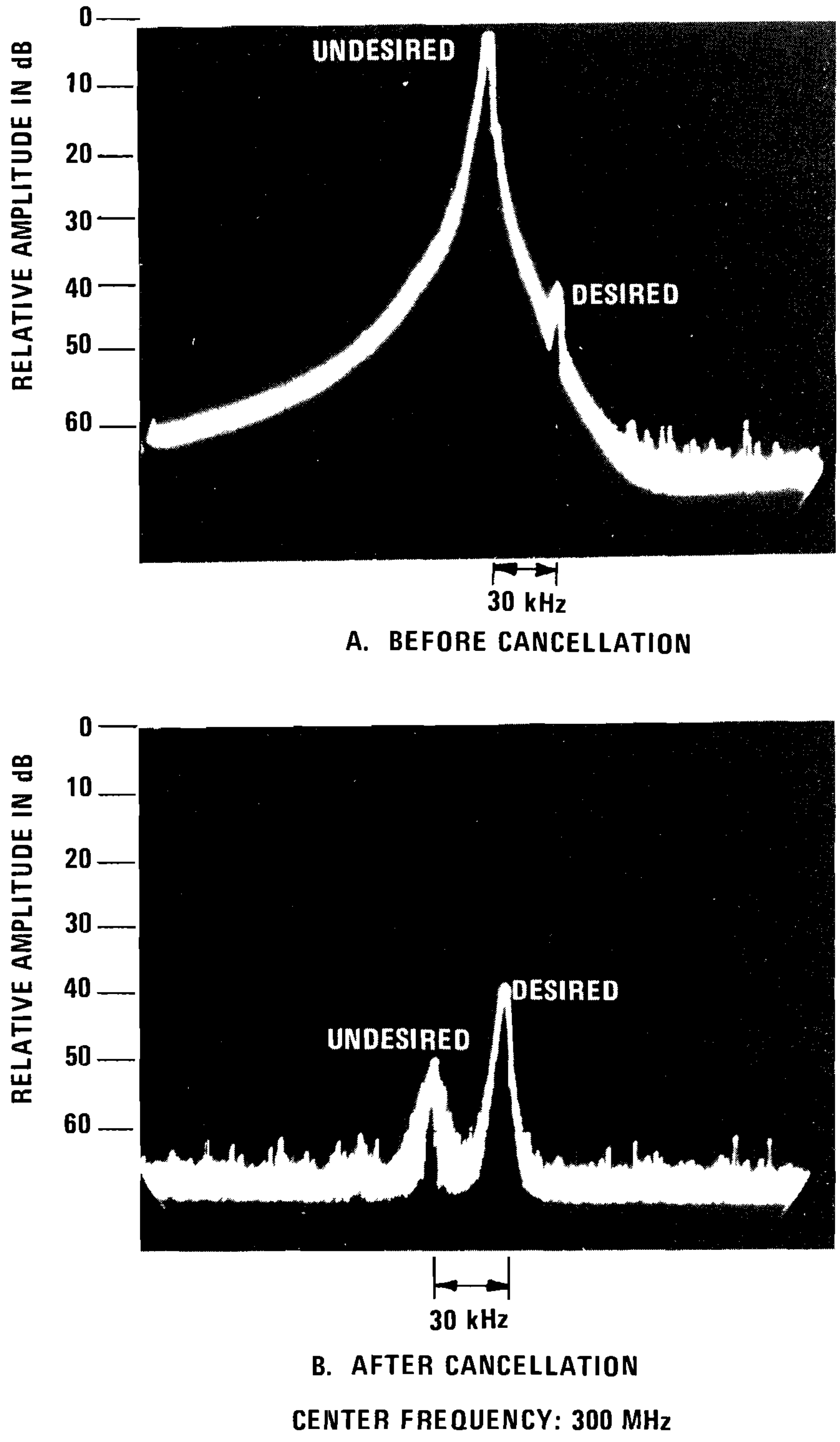


Figure 21. Spectrum Analyzers Displays Which Show the CW Suppression Capabilities of the AM Cancellation Filter.

SECTION IV

BROADBAND LINEAR AMPLIFIERS

A. General Considerations

The extension of the dynamic range of active filters is closely related to an improvement in their linearity characteristics and power handling capabilities. The maximum power that can be handled is determined by the dissipation characteristics of the active device employed. If distortion in the signal path is to be minimized, the device must operate in a linear mode well below its usual saturation power level. Hence, the maximum power level that the filter can handle is determined by the dissipation and saturation characteristics of the active elements. The total dynamic range is the total voltage swing that the device can handle between its saturation level and its residual noise level. The dynamic range can be extended by reducing the residual noise level and raising the saturated power level. The residual noise level can be lowered by the use of active elements of lower noise figures in circuit configurations conducive to low noise operation. A straightforward approach to increasing the saturated output power level is to utilize active elements capable of high power dissipation. Recent advances in solid state technology have resulted in transistors capable of delivering several watts in the UHF region.

Figure 22 illustrates the manner in which the actual transfer characteristic of a typical active element deviates from the ideal behavior. The characteristics of a device with improved linearity is also illustrated. In the region between the residual noise level and the saturation level of the device, the transfer characteristic will exhibit some deviation from a straight-line behavior. This nonlinear behavior determines the performance of the device in a multiple signal environment. Cross modulation, intermodulation and waveform distortion are produced by the nonlinearities. Consequently, it is desirable to maximize the linearity of the transfer function over as wide a dynamic range as possible.

For the ideal transfer characteristic shown in Figure 22, the output voltage is a constant multiple of the input voltage, that is

$$e_o = a_1 e \quad . \tag{14}$$

For the practical case, however, the transfer characteristic is not linear, and is more appropriately represented by

$$e_o = a_1 e + \sum_{n=2}^{\infty} a_n e^n \quad , \tag{15}$$

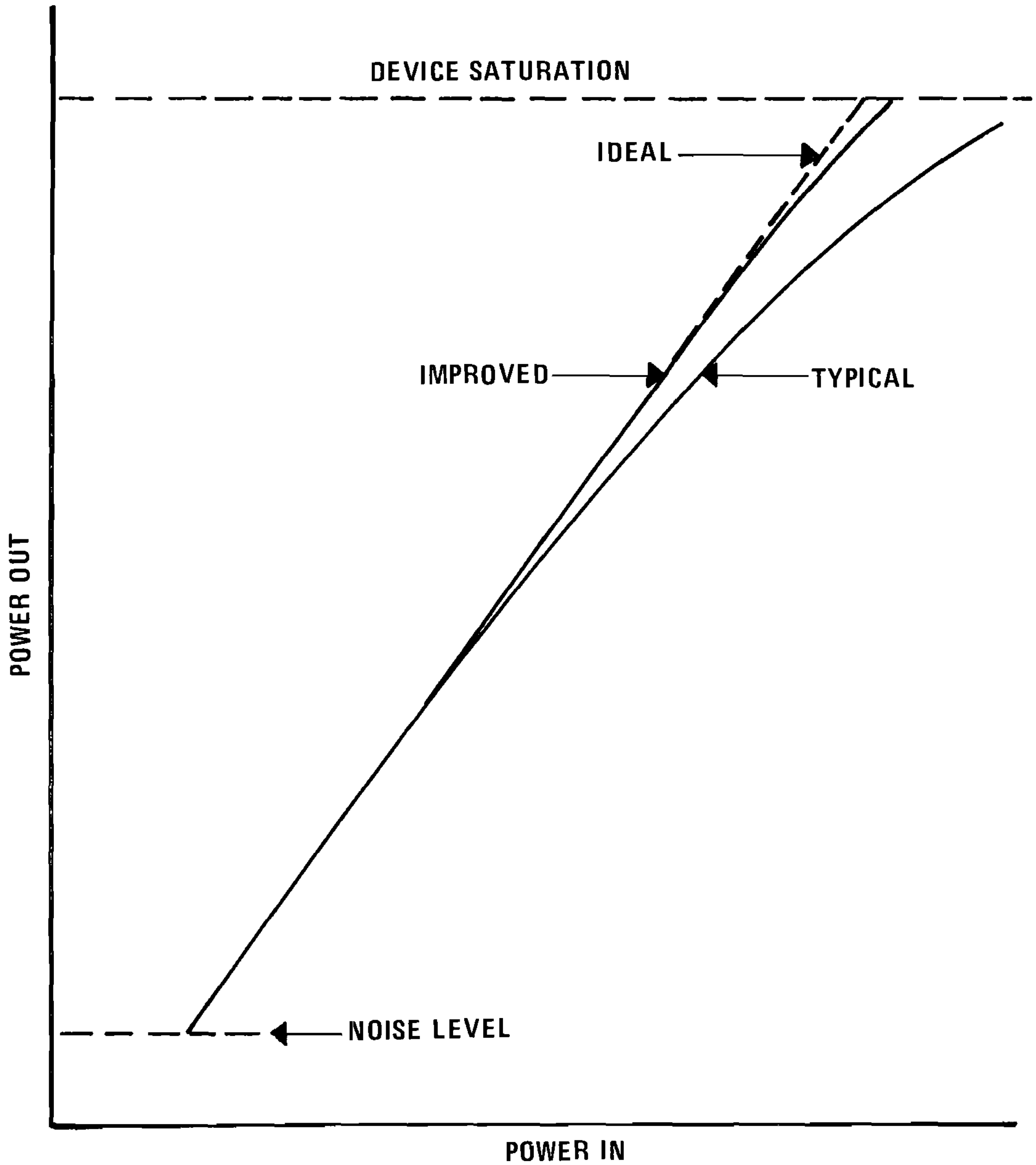


Figure 22. Comparison of Typical Amplifier Transfer Characteristics.

where e = the effective input voltage,

e_o = the output voltage, and

a_n = power series coefficients whose values are dependent upon the characteristics of the nonlinearity.

Thus, in addition to the desired output, a_1e, distortion products will be present caused by the higher order terms a_2e^2, a_3e^3, etc.

Negative feedback has long been recognized as an effective technique for reducing nonlinear distortion products in amplifier circuits. Consider the feedback network of Figure 23, where e is the applied voltage. The effective input voltage is

$$e_i = e - \beta e_o \quad , \tag{16}$$

assuming negative feedback, β being the fraction of the output voltage that is fed back to the input.

Substituting (15) into (16):

$$e_i = e - \beta a_1 e_i - \beta a_2 e_i^2 - \beta a_3 e_i^3 - \dots \quad , \tag{17}$$

or

$$e_i + \beta a_1 e_i = e - \beta a_2 e_i^2 - \beta a_3 e_i^3 - \dots \quad . \tag{18}$$

Then,

$$e_i = \frac{e}{1 + a_1\beta} - \frac{a_2\beta e_i^2}{1 + a_1\beta} - \frac{a_3\beta e_i^3}{1 + a_1\beta} - \dots \quad . \tag{19}$$

Panter [10] provides a technique for obtaining the output voltage e_o in terms of the applied voltage e. From the expression for e_i in (19), he proceeds to obtain the basic feedback equation

$$e_o = a_1 e + \frac{a_2 e^2}{1 + a_1\beta} + \left[\frac{a_3}{1 + a_1\beta} - \frac{2a_2^2\beta}{(1 + a_1\beta)^2}\right] e^3 + \dots \quad , \tag{20}$$

where the input voltage e has been increased by the factor $(1 + a_1\beta)$ to restore the output to its level without feedback.

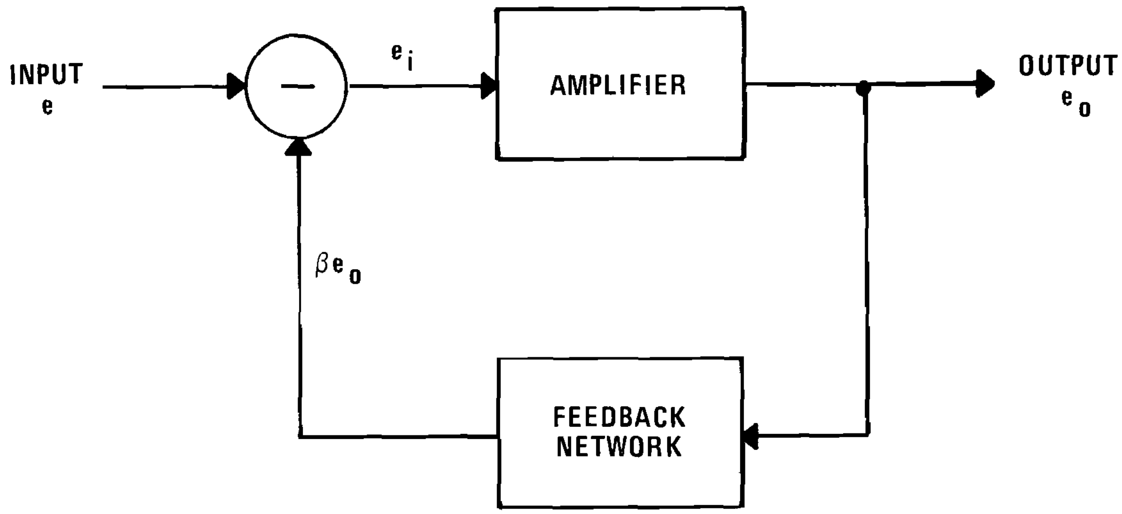


Figure 23. Diagram of a Basic Negative Feedback System.

The form of equation (20) makes it difficult to evaluate the merits of negative feedback in reducing distortion products. The equation is usually approximated by assuming that $a_1\beta > 1$ and that the coefficient a_1 is much greater than the coefficients of the higher order terms. Under these assumptions, equation (20) can be expressed as

$$e_o = a_1 e + \frac{a_2}{1 + a_1\beta} e^2 + \frac{a_3}{1 + a_1\beta} e^3 + \ldots \quad , \tag{21}$$

or

$$e_o = a_1 e + \frac{1}{1 + a_1\beta} \sum_{n=2}^{\infty} a_n e^n \quad . \tag{22}$$

Equation (22) shows that the coefficients of the higher order terms in the nonlinear transfer characteristic have been reduced in amplitude by the factor $\frac{1}{1 + a_1\beta}$; hence harmonics or distortion products generated by these higher order terms will be reduced by this factor.

Another method of improving amplifier performance is through the use of power division. Rather than using a single amplifier, suppose the signal to be amplified is split into k paths with a k-way power divider as shown in Figure 24. Each of the k paths contains an amplifier to provide the desired gain. The k outputs of the amplifiers are then combined using another k-way signal divider.

Several advantages result from the power division technique. First, a high level signal may be divided into a number of lower level signals which can be handled with amplifiers of lower power rating. Second, this technique allows the voltage swing at the input to the amplifiers to be restricted to a more linear portion of the transfer characteristics.

The effect of power division in reducing distortion products can be illustrated from equation (15), which is repeated below.

$$e_o = a_1 e + \sum_{n=2}^{\infty} a_n e^n \quad . \tag{23}$$

Suppose that the input signal, e, is divided into k paths, each containing an amplifier with a transfer characteristic represented by equation (23). The signal power in each path would thus be proportional to e^2/k and each amplifier output would be represented by

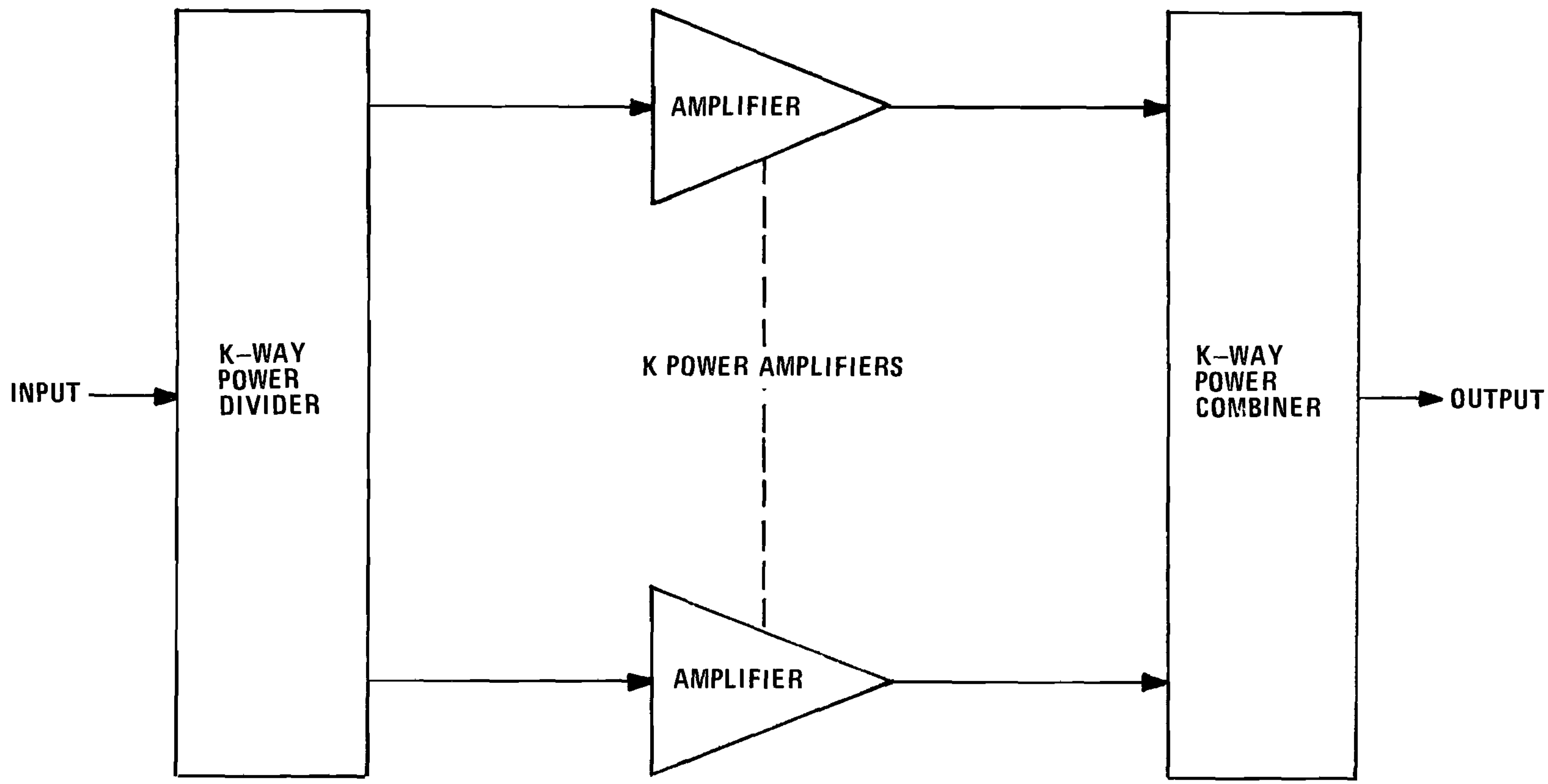


Figure 24. Illustration of Power Sharing Mode of Operation.

$$e_{ok} = a_1 \frac{e}{\sqrt{k}} + \sum_{n=2}^{\infty} a_n \left(\frac{e}{\sqrt{k}}\right)^n \quad . \qquad (24)$$

To recover the desired signal, the signal power in each individual path must be summed. But the power in each path is proportional to

$$P_{ok} \propto (e_{ok})^2 = \left[a_1 \frac{e}{\sqrt{k}} + \sum_{n=2}^{\infty} a_n \left(\frac{e}{\sqrt{k}}\right)^n\right]^2 \quad . \qquad (25)$$

Summing for k paths,

$$\sum^{k} P_{ok} = k \left[a_1 \frac{e}{\sqrt{k}} + \sum_{n=2}^{\infty} a_n \left(\frac{e}{\sqrt{k}}\right)^n\right]^2 \quad , \qquad (26)$$

and the output voltage would thus be

$$e_o = \sqrt{k} \left[a_1 \frac{e}{\sqrt{k}} + \sum_{n=2}^{\infty} a_n \left(\frac{e}{\sqrt{k}}\right)^n\right] \quad . \qquad (27)$$

or

$$e_o = a_1 e + \sqrt{k} \sum_{n=2}^{\infty} a_n \left(\frac{e}{\sqrt{k}}\right)^n \quad . \qquad (28)$$

Equation (28) may also be expressed as

$$e_o = a_1 e + \frac{a_2 e^2}{(k)^{\frac{1}{2}}} + \frac{a_3 e^3}{k} + \frac{a_4 e^4}{(k)^{3/2}} + \ldots + \frac{a_n e^n}{(k)^{\frac{n-1}{2}}} \quad , \qquad (29)$$

which illustrates the reduction in amplitude that will occur for distortion products generated by the higher order terms.

As an example of harmonic reduction that can be achieved through power division, consider the case where the amplifier input signal is

$$e = V \cos \omega t \quad . \tag{30}$$

Expanding this input in equation (29) yields harmonic terms of the form $C_n \cos n\omega t$, where the harmonic coefficient, C_n, is in series form. If C_n is approximated by neglecting contributions from higher order terms in the expansion (i.e., if C_n is derived only from the lowest order term in equation (29) which generates the nth harmonic), the amplifier output from equation (29) is

$$e_o = a_1 V \cos \omega t + \frac{a_2 V^2}{2(k)^{\frac{1}{2}}} \cos 2\omega t + \frac{a_3 V^3}{4k} \cos 3\omega t + \frac{a_4 V^4}{8(k)^{3/2}} \cos 4\omega t + \ldots \tag{31}$$

Without power division, the output would be represented by an expansion of equation (30) in equation (15), which would yield

$$e_o = a_1 V \cos \omega t + \frac{a_2 V^2}{2} \cos 2\omega t + \frac{a_3 V^3}{4} \cos 3\omega t + \frac{a_4 V^4}{8} \cos 4\omega t + \ldots \quad . \tag{32}$$

A comparison of equations (31) and (32) reveals that the use of power division reduces the second harmonic voltage by $1/\sqrt{k}$, the third harmonic voltage by $1/k$, etc. In general, the nth harmonic voltage will be reduced by the factor $\left[\frac{1}{[k]^{\frac{n-1}{2}}}\right]$, where $n \geq 2$. The nth harmonic power would thus be reduced by $\left[\frac{1}{[k]^{(n-1)}}\right]$.

B. Experimental Results

The effectiveness of negative feedback in reducing distortion is shown by equation (20). Equation (19) shows, however, that reduced fundamental gain must be tolerated to obtain reduction of distortion products. For example, broadband negative feedback is often realized with an unbypassed series emitter resistance. Since this resistance is a part of the ac path, it directly lowers the gain of the circuit.

A common emitter configuration utilizing a 2N3866 transistor was evaluated for harmonic generation as a function of power output level with two different values of emitter resistance. Figure 25 shows that the level of the second harmonic with a 47 ohm resistance was typically 5 dB lower than the level with a 10 ohm resistance. Below saturation, the third harmonic levels fall closely together for both resistance values. With the 10 ohm resistance, the third harmonic levels show a dramatic upswing above +20 dBm.

Since the common base configuration is often recommended as a broadband amplifier in the VHF and UHF ranges, a common base 2N3866 amplifier was evaluated for its harmonic generation characteristics. The curves shown in Figure 26 illustrate that high harmonic levels were observed in the common base amplifier. Typically the second harmonic levels appear to be approximately 30 dB higher in the common base amplifier than in the common emitter amplifier. The erratic nature of the third harmonic variation makes a relative comparison difficult; however, over the range of power levels tested, the common emitter amplifier demonstrates lower third harmonic generation. On the basis of these measurements, further development emphasized the use of the common emitter configuration.

The 2N3866 represents a class of transistors whose collector is tied directly to the case. In any configuration except common collector, the case must be left floating, i.e., the collector junction is effectively RF exposed and is very susceptible to circuit parasitics. In the common emitter amplifier, the case cannot be thermally connected to the chassis without destroying the frequency response due to the shunt capacity of the heat sink. Consequently, high power operation is restricted because of the limited heat dissipation available with the 2N3866 transistor.

Good thermal conductivity with electrical isolation between terminals and case is obtained with stud-mounted transistors such as the 2N5090 and 2N3375. Both transistors exhibit RF characteristics comparable to the 2N3866 and are capable of dissipating greater than 5 watts. Although the published data indicate that the two transistors are quite similar, experimentally, the 2N3375 exhibited less stability and required careful adjustment of the dc bias to obtain stable operation. Several amplifiers were constructed with both the 2N5090 and 2N3375 and, overall, the 2N5090 exhibited the more desirable characteristics of improved stability, greater bandwidth, and less complex tuning for a flat frequency response.

The circuit configuration shown in Figure 27 is a good basic single-stage amplifier in that it exhibits a flat gain response as shown by Figure 28, shows relatively low pertubation of the gain characteristics, and maintains stability when cascaded with other similar stages. Two of the basic amplifiers were combined in the push-pull configuration shown in Figure 29 to produce a building-block amplifier which was used to evaluate the effectiveness of power sharing techniques to obtain improved linearity. Two of these building-block amplifiers were combined in the

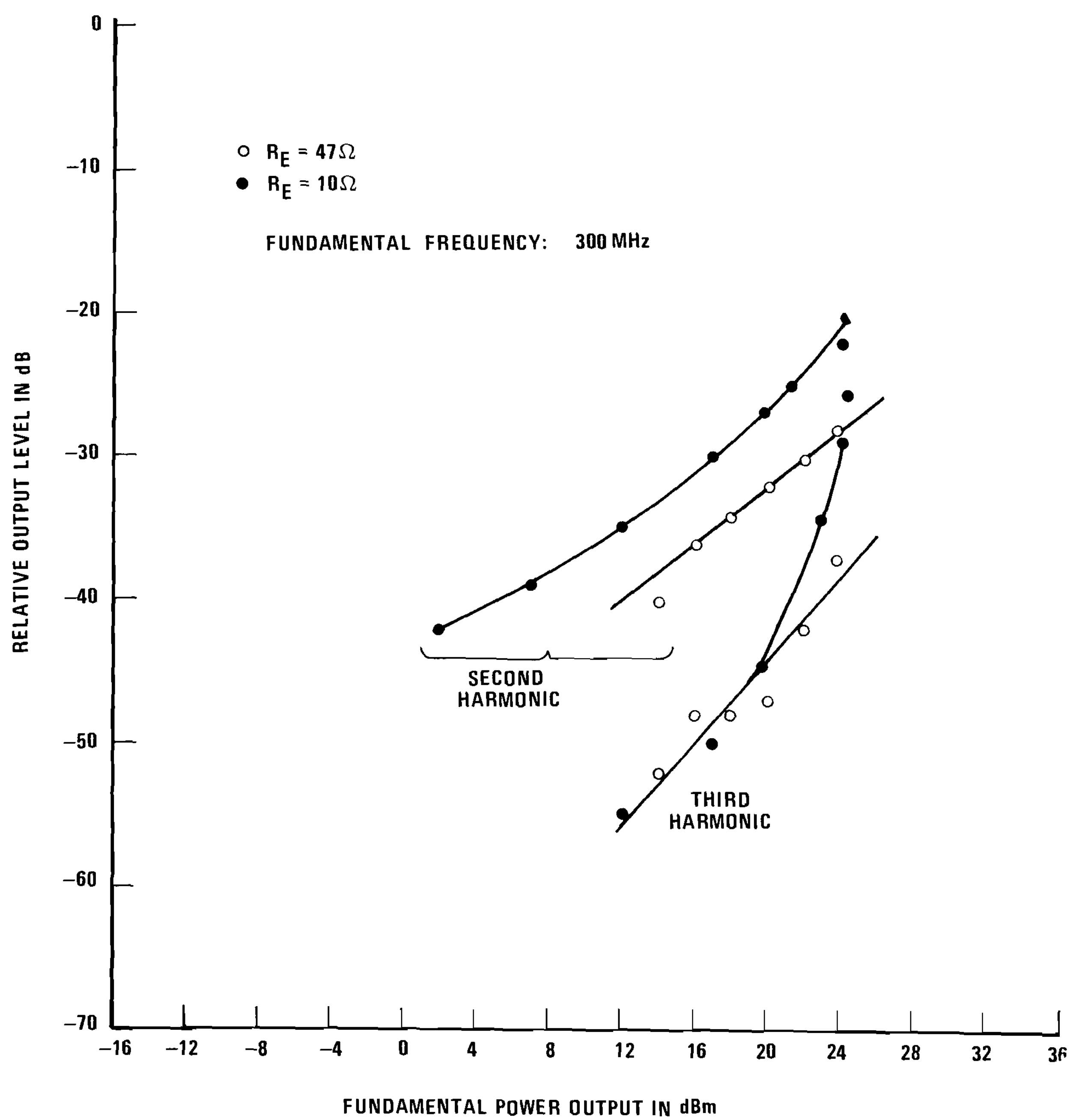


Figure 25. Harmonic Generation Levels for Two Values of Emitter Resistances.

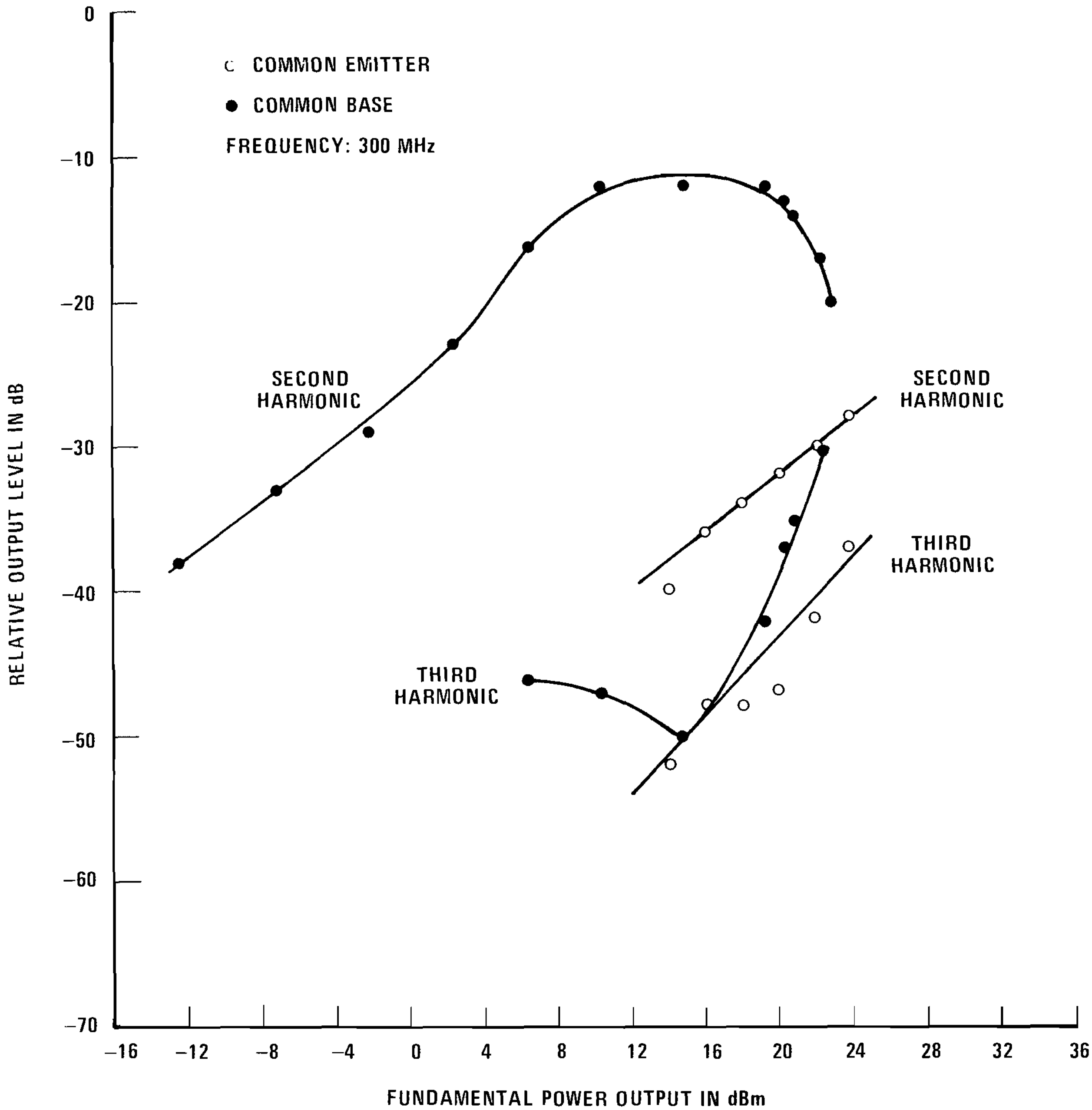
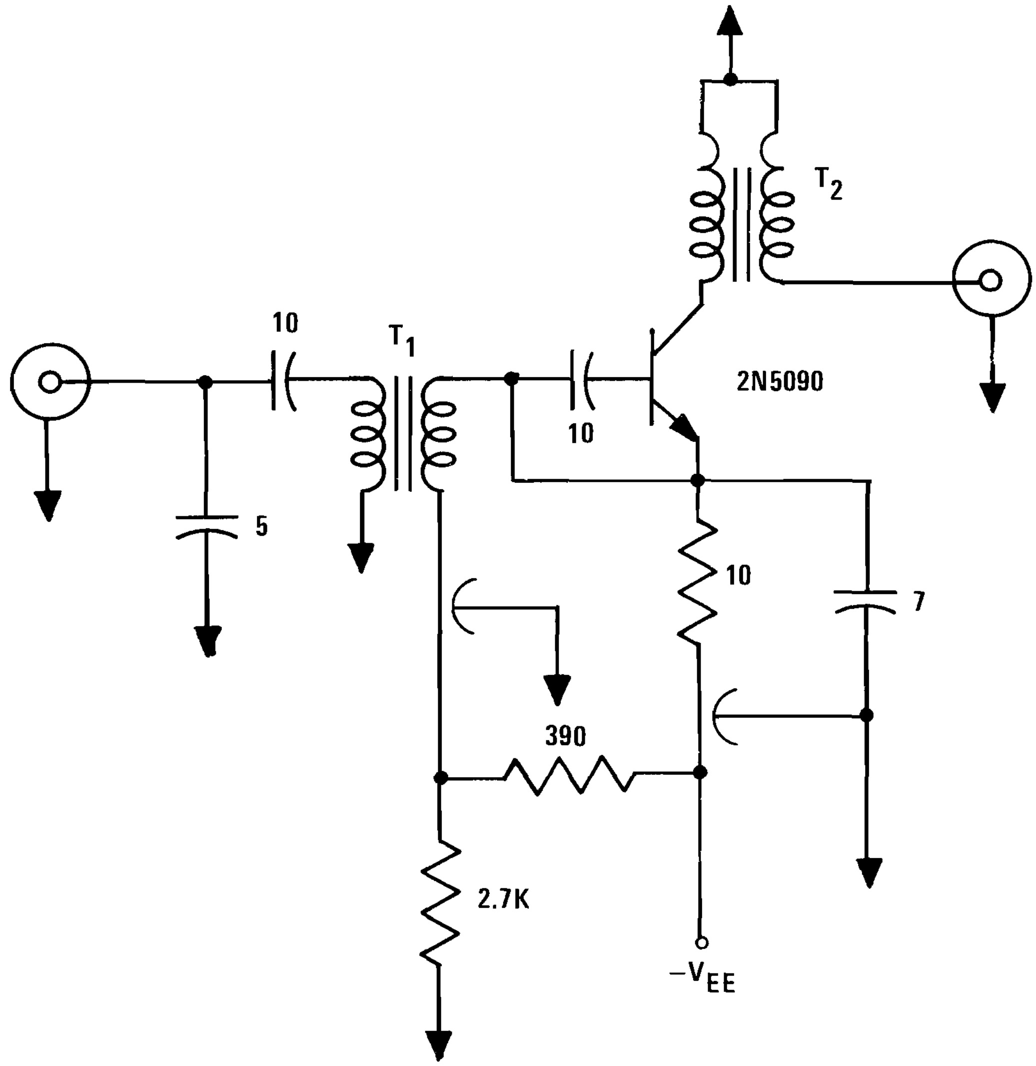


Figure 26. A Comparison of the Harmonic Levels Generated by a Common Base Amplifier and by a Common Emitter Amplifier.



T_1, T_2: 1:1 BROADBAND TRANSFORMERS

Figure 27. Schematic Diagram of a Single Stage Broadband Amplifier.

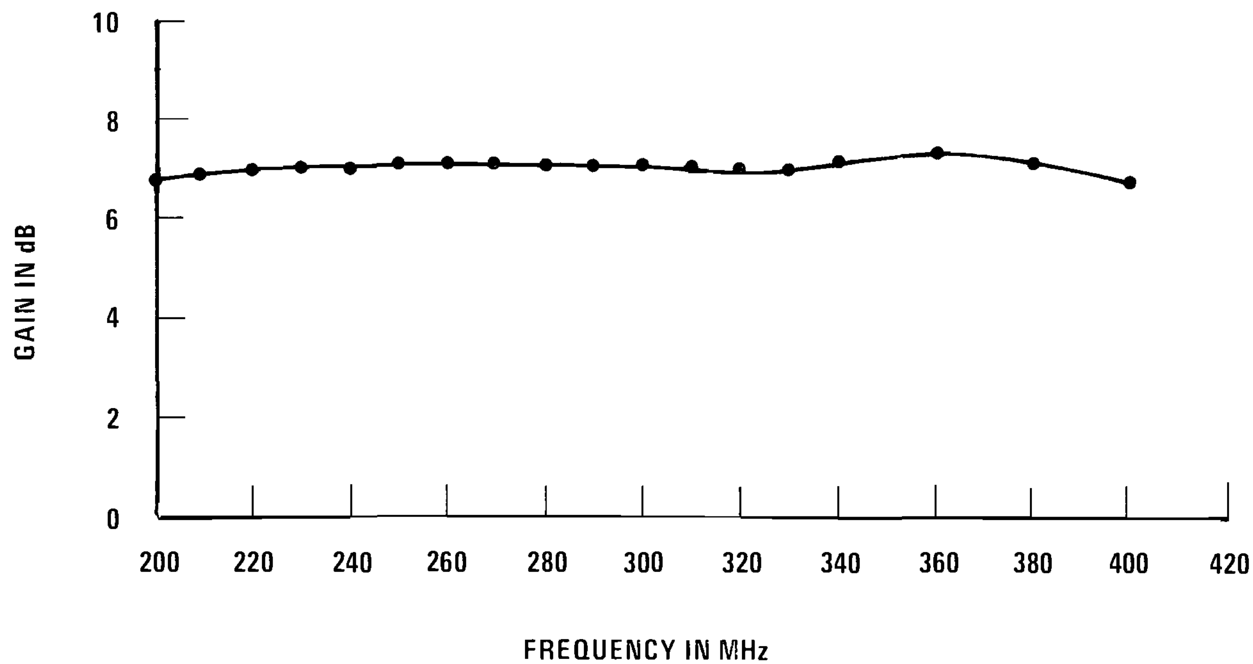


Figure 28. Gain Characteristics of the Single Stage Amplifier.

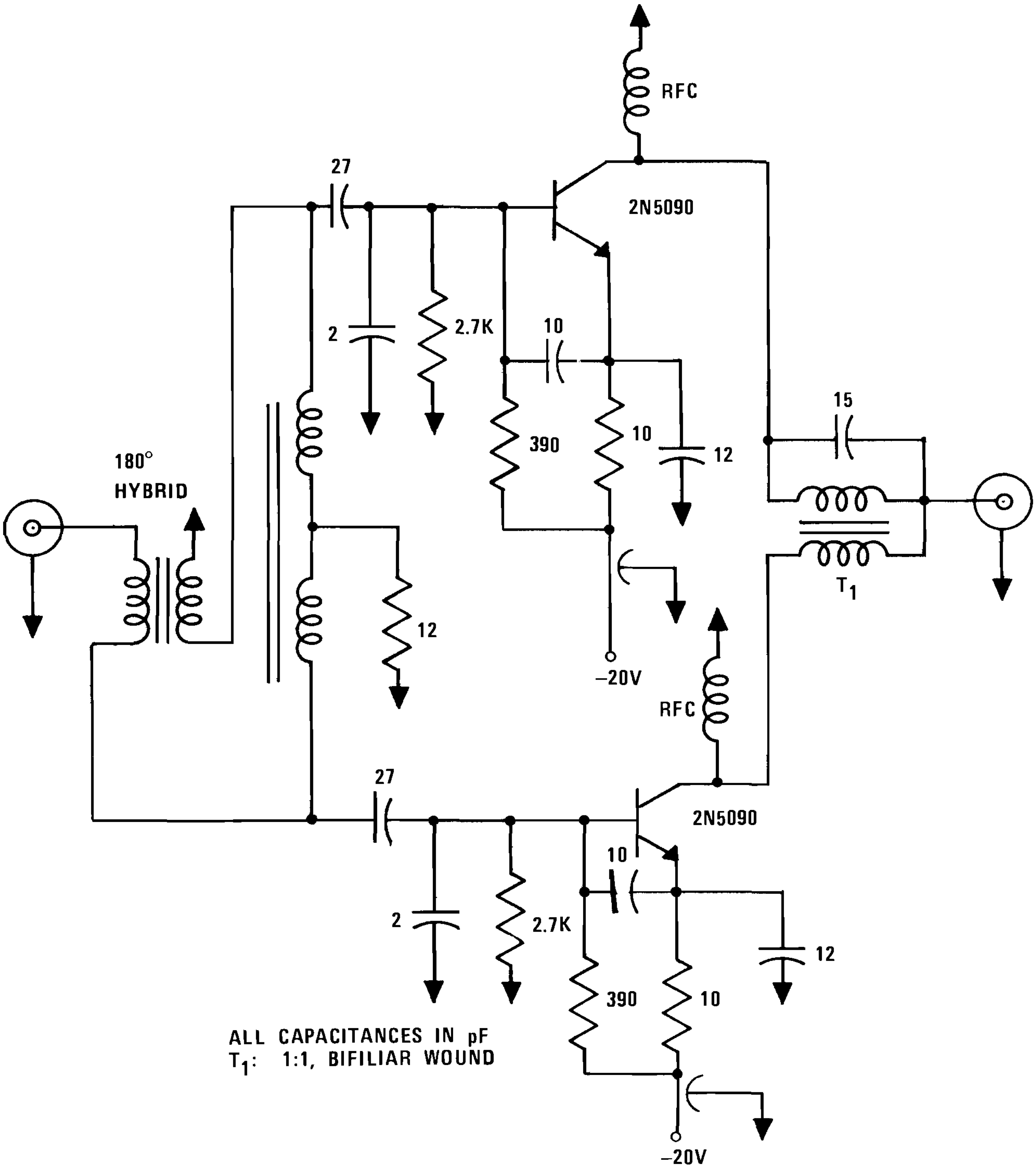


Figure 29. Basic Building-Block Amplifier Consisting of Two Parallel Stages.

push-pull arrangement shown in Figure 30 which was in turn combined in parallel in another similar arrangement to produce an amplifier consisting of eight basic transistor amplifiers in parallel.

In the basic amplifier, a compromise between improved linearity and reduced gain is achieved by using a 10 ohm unbypassed emitter resistor to supply broadband negative feedback. With this resistance, the basic amplifier typically exhibited 6 dB gain over the 200 to 400 MHz range.

Parallel amplifiers can be operated either in the push-pull mode or in the push-push mode. For push-pull operation the amplifiers are driven out-of-phase and summed out-of-phase whereas in the push-push mode of driving signals are in phase and summation is provided by the in-phase ports of the hybrid. The relative levels of the second harmonic generated in each mode are illustrated in Figure 31 in comparison with the levels generated in a single stage amplifier. Push-push operation of paralleled amplifiers achieves a 6 dB reduction in the second harmonic level over the single amplifier. Push-pull operation provides an additional 5 dB of second harmonic reduction for output levels less than +20 dBm. The third harmonic level was about the same for either parallel mode.

The levels of second and third harmonic generation for a single stage amplifier and for two, four and eight push-pull stages are shown in Figure 32. The level of second harmonic generation for four stages in parallel is approximately 10 dB less than the levels generated in a single stage amplifier. Eight parallel stages exhibit an additional 10 dB reduction. Approximately 20 dB reduction in the third harmonic level is achieved in going from a single stage amplifier to eight parallel stages. The eight-stage amplifier shows a second harmonic level of -42 dB and a third harmonic level of -58 dB at a one-watt output level.

This amplifier development program has shown the improvement in linearity which may be expected from the application of negative feedback and power sharing techniques to UHF, broadband transistor power amplifiers. Negative feedback in the nature of unbypassed emitter resistance is quite effective; however, a compromise between linearity improvement and gain reduction must be made. As expected, significant reduction in the level of harmonics is achieved by the use of parallel broadband stages. Both second and third harmonic levels are reduced by parallel operation. The data also indicate that the push-pull mode is preferred because this mode provides greater rejection to second harmonic generation than the push-push mode. As expected, these linear amplifiers are not very efficient. For example, to provide the one-watt fundamental output with the second harmonic down 42 dB and the third harmonic down 58 dB, ten watts of dc power were required for an efficiency of only 10 per cent.

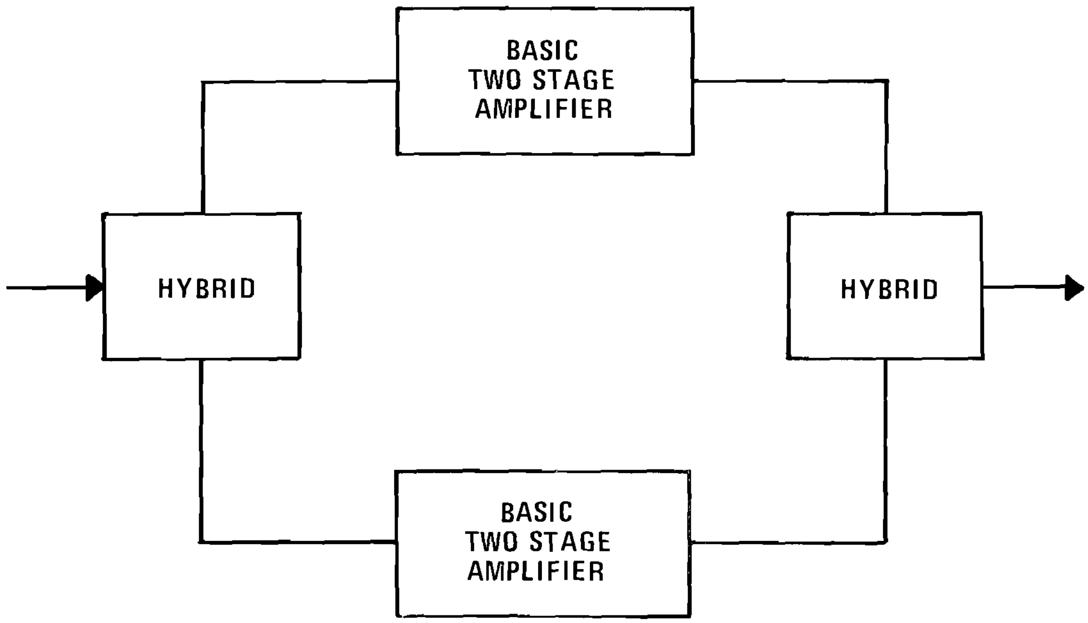


Figure 30. Block Diagram of Paralleled Amplifier Arrangement for Push-Pull or Push-Push Operation.

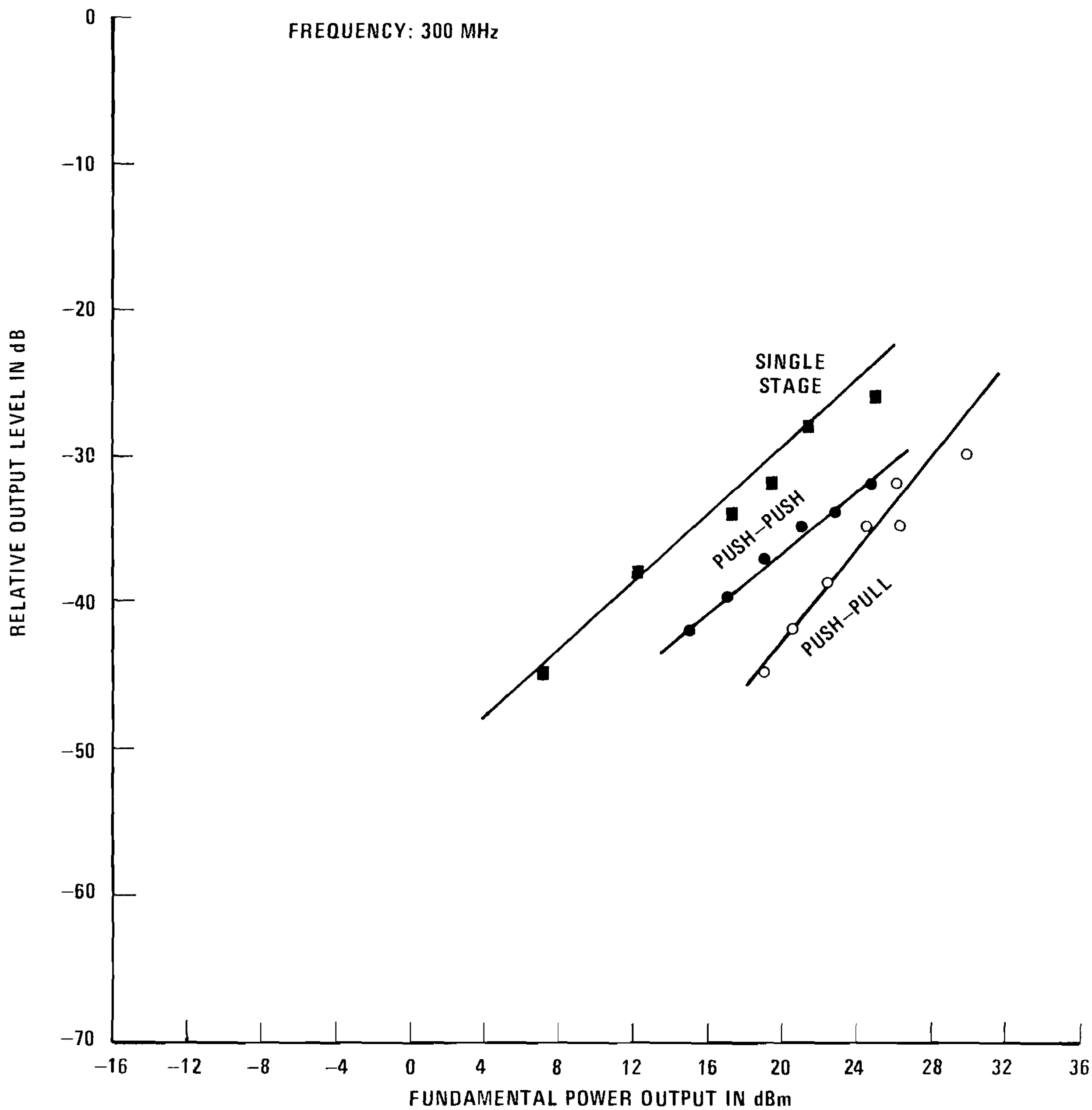


Figure 31. Second Harmonic Generation Characteristics of One-Stage, Two-Stage and Four-Stage Amplifiers.

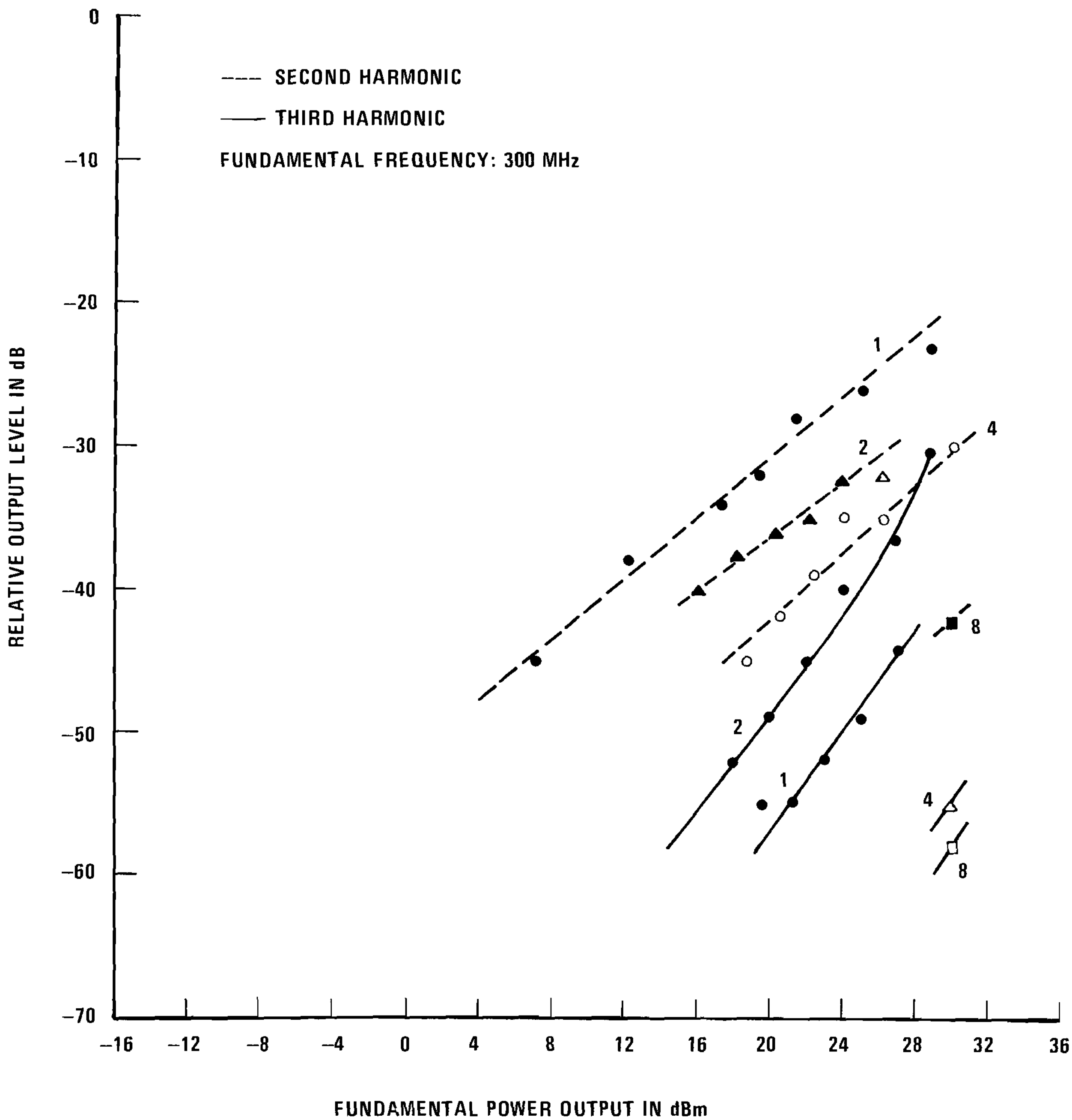


Figure 32. Harmonic Generation Characteristics of Various Amplifier Configurations.

SECTION V

CONCLUSIONS

The instabilities normally associated with high orders of multiplication can be avoided through the use of multiple feedback loops. In this way, extremely high Q's can be obtained from small sized resonators of moderate Q. Although active Q multiplication improves the relative stop band attenuation of the resonator, the rate of attenuation rolloff is still determined by the number of resonator stages. If steep skirt characteristics are required, multiple stages in series are necessary. The advantages presented by cascaded stages of Q multiplication are a narrow bandwidth combined with a high skirt selectivity.

It was shown analytically and experimentally that a negative resistance function which is effective over a broad range of frequencies can be realized. This negative resistance can be used to enhance the Q of coaxial cavities and other passive resonators in the 200 to 400 MHz region. To assure stability, the multiplication factors obtainable with the technique are limited to 10 or less. However, the technique is operationally simple in that relatively non-critical tuning adjustments are all that are required for operation over a wide frequency range. In view of the promise of the negative resistance technique, further investigation into the integration of negative resistance amplifiers directly into resonators should be pursued.

The performance of an active filter in a multiple signal environment is directly affected by the dynamic range and by the linearity of the transfer function within the limits of the dynamic range of the amplifiers in the filter. Negative feedback and power sharing techniques were effectively applied to broadband transistor amplifiers to achieve one watt output capabilities with second and third harmonic levels down at least 40 dB and 55 dB, respectively. As is generally true with any active device operating in a highly linear mode, the efficiency of the amplifiers is low, being in the neighborhood of 10 per cent.

The feasibility of suppressing closely spaced CW and AM interference through signal synthesis was shown with the breadboard model of the AM cancellation filter. Although the filter possesses certain limitations, it promises to fill a definite need for a suppression device in operational situations which do not permit the acquisition of a sample of the culprit signal directly from its source.

SECTION VI

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13. ABSTRACT

Several active filter techniques for the reduction of receiver interference in the 225 to 400 MHz range are described. Positive feedback Q multiplier techniques were extended to include (1) the use of multiple feedback loops to achieve a high order of stable multiplication in each stage and (2) the use of cascaded stages of Q-multiplied resonators to obtain improved skirt selectivity. Negative resistance Q multiplication was achieved over a wide frequency range through the development of a common collector transistor amplifier that exhibits stable negative resistance properties in the UHF region. The negative resistance amplifier was incorporated into a breadboard model of a tunable filter which employs both active and passive stages to produce a high Q response characteristic with high skirt selectivity over the entire band. An AM cancellation filter that achieves suppression of an unwanted signal by cancellation via a synthesized replica of the signal was developed. The breadboard model demonstrated a suppression capability of 30-35 dB for AM signals and about 50 dB for CW signals.

To enhance the capabilities and versatility of UHF active interference suppression filters, linearization techniques for broadband solid state amplifiers were investigated. The application of negative feedback and the use of the push-pull mode of parallel operation provided a significant reduction in harmonic generation while retaining good gain-band-width characteristics in amplifiers of one-watt power output capabilities.

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14. KEY WORDS	LINK A		LINK B		LINK C	
	ROLE	WT	ROLE	WT	ROLE	WT
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Active Filters						
Q Multiplication						
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