

Vol. 6 No. 1 January/February 1979

# Cascadable Amplifiers

WATKINS-JOHNSON COMPANY

# Tech-notes

## CASCADABLE AMPLIFIERS

TO-8 cascable amplifiers are of miniature size, have a high degree of circuit flexibility, and are used in sophisticated navigation and guidance systems; radar and sonar equipment; ground based, airborne and satellite communications equipment; telemetry electronics; electronic countermeasures systems and in electronic warfare equipment. In short, TO-8 cascable amplifiers are used in virtually any high-quality RF and IF amplifying or processing application that covers their respective frequency ranges.

Since these amplifiers are designed to operate in a tandem configuration, they are referred to as being "cascable." By definition, this means that the amplifiers' input and output VSWR and stability factors are such that the units can be cascaded to obtain higher gain and, in addition, a power unit can be used as output stage and a low noise unit as input stage. Figure 1 shows a typical four-stage cascade with a low noise input block, a gain stage, a driver stage, and a power output stage. Cascability implies that the resultant VSWR on overall input and output ports remains acceptable for the purpose of overall system operation (typically  $< 2.0:1$ ), and that the gain flatness of the unit is not appreciably altered. In addition,

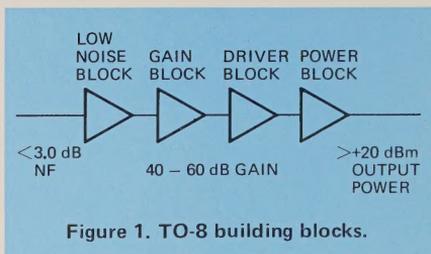


Figure 1. TO-8 building blocks.

cascability also implies that the units remain unconditionally stable.

The purpose of this article is to discuss the basic circuit concept behind the cascable amplifier and to detail some of the techniques that are used to maintain hardware stability in the application of the units.\* The fundamental common denominator to most broadband cascable amplifiers is collector-to-base, shunt-resistive feedback and series-emitter resistive feedback, as shown in Figure 2. There are some reactive feedback methods that have been employed both in combination with and without resistive feedback to reduce the degradation that the resistive feedback has on the noise figure. However, the consideration of these design schemes is beyond the scope of this article.

The typical transducers used in the TO-8's are microwave bipolar transistors which have a nominal 6 dB/

### \* HISTORICAL NOTE

Combined plate-to-grid and cathode feedback was shown in 1951 by Richard S. Burwen for a vacuum tube triode. Howard Lefkowitz later applied collector-to-base and series-emitter feedback simultaneously to a bipolar transistor, and showed the two-port representation for computer analysis.

More recently, combined drain-to-gate and series-source feedback resistances have been successfully applied to both V-FET's and GaAs FET's, indicating that the three major types of electronic transducers can have their gain and impedance levels effectively altered for broadband operation without the use of passive gain tailoring networks.

### References

1. Burwen, R.S. "Equivalent Circuits to Simplify Feedback Design," *Audio Engineering*, (October 1951), pp 11, 12, 42-45.
2. Lefkowitz, H. "Transistor AC Amplifier Uses Multiple Feedback," *Electronics*, (May 23, 1958), pp 84, 85.

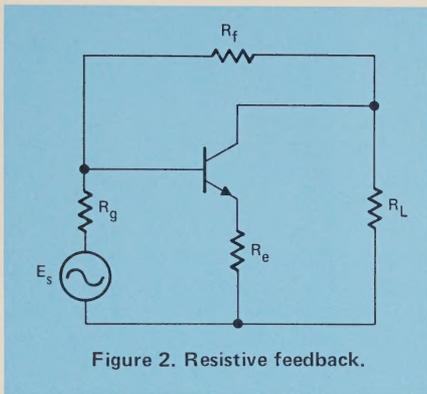


Figure 2. Resistive feedback.

octave gain slope, as shown in Figure 3. The amplifier gain curve can be flattened using pure resistive feedback with no reactive elements up to a frequency where the open loop gain is approximately 10 dB greater than the closed loop gain. Then, reactive elements must be used to provide input and output match and to “gain-peak” the amplifier to obtain the highest gain bandwidth product. A typical RF circuit excluding DC biasing, coupling, and bypass capacitors is shown in Figure 4. The parasitic

elements, such as header capacitance and bondwire inductance, are absorbed into the circuit elements. Manual computer optimization is used on a specially developed W-J computer program that is specifically designed for analyzing cascaded amplifier circuits.

This program uses conventional matrix techniques to treat each circuit element and uses fixed topology. Single-stage gain typically runs from 6 to 20 dB, with  $R_e$  ranging from 15 to 0 ohms and  $R_f$  ranging from 150 to 800 ohms, as illustrated in Figure 5. For low noise circuits,  $R_e$  usually goes to zero, since the transistor is normally biased for less than 10 mA collector current and the internal  $R_e$ , which can be approximated by  $.026/I_C$ , acts as sufficient emitter feedback. The curves in Figure 6 show the gain, and input and output impedance as a function of  $R_e$  with a fixed  $R_f$ .  $R_e$  is fixed and  $R_f$  varied in Figure 7, which shows the inverse gain relationship between these two resistors. It is this property of the microwave bipolar transistor that allows various gain options to be realized. For a fixed value of  $R_e$ , to obtain as close to a 50-ohm input and output impedance as possible, a

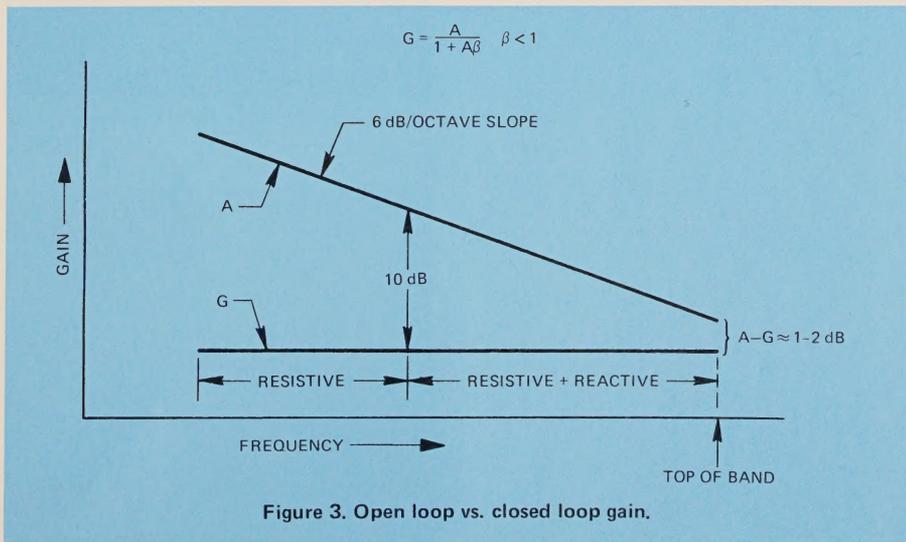


Figure 3. Open loop vs. closed loop gain.

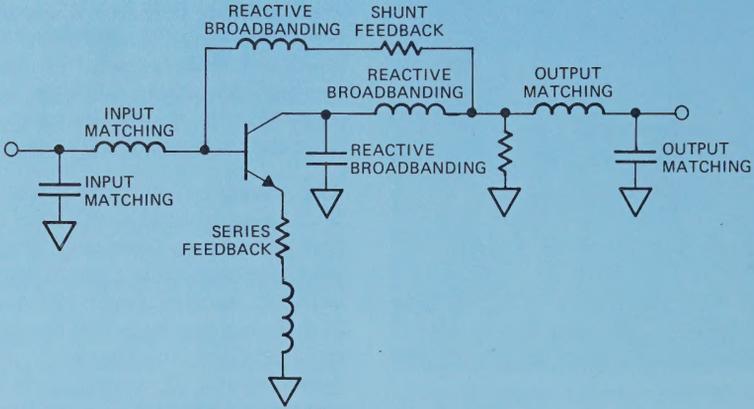


Figure 4. Typical RF circuit.

certain value of  $R_f$  which will dictate the gain is required. For higher gain, a lower value of  $R_e$  is chosen, which requires an increase in the value of  $R_f$ . The gain that can be obtained at the top of the band is typically within 2 dB of the transducer's open-loop gain. To obtain more than this gain

from the device usually results in both lack of unconditional stability and increased input VSWR.

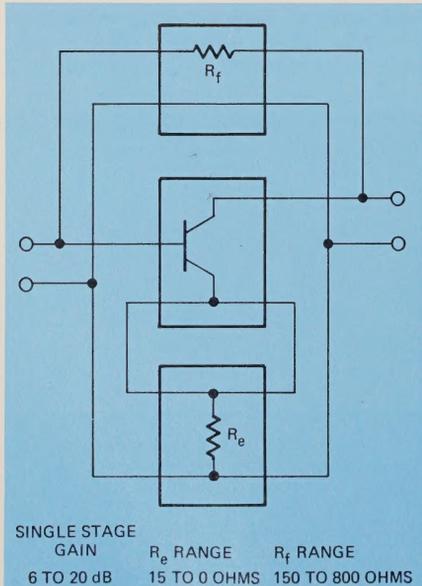


Figure 5. Two-port topology.

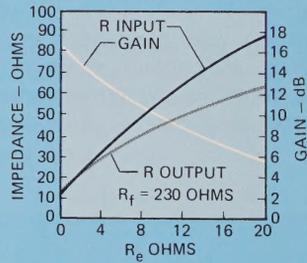


Figure 6. Gain and impedance.

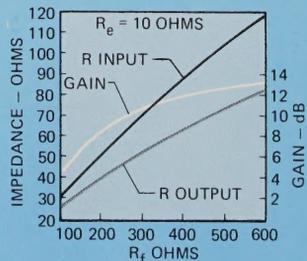


Figure 7. Gain and impedance vs. collector-base resistance.

The circuit design of the cascaded amplifier is broken up into two distinct steps: 1) the low-frequency or resistive portion, and 2) the high-frequency or reactive element portion. This can be done manually as opposed to using automatic optimization, so that the designer maintains full control of the changes in each element value, thereby maintaining control in the overall design process. By following this optimization procedure, the designer learns what the various sensitivities of each element are and, as a result, develops greater insight into how the elements interact with each other. Automatic optimization has the advantage, however, of requiring less skill in the final determination of the optimized element values, but the procedure must be exercised with caution.

An example of the type of problem that can occur when using automatic optimization is shown in Figure 8. The initial element values used were determined by first performing an initial manual run on a 100-2600 MHz

single-stage amplifier. The first automatic trial used four frequencies and equal weights on S11, S22, and S21 (input VSWR, output VSWR, and gain). The midband VSWR at 1.5 GHz improved from 1.85:1 to 1.45:1, but at 2.6 GHz it degraded from 1.8:1 to 3.0:1. Increasing the weighting factors of S11 and S22 by 3 over that of S21 and reducing the number of frequencies, decreased the output VSWR at 2.6 GHz to 2.14:1, which was still higher than the initial manual result. Further reduction could be made, but it would probably require an even further reduction in the number of frequencies used, thus moving the method closer to the manual technique.

### Noise-Figure Degradation

For a rough first approximation of the quantitative degradation in noise figure that the resistive feedback elements cause, the following derivations may be used. The approximation excludes the actual transistor and its noise properties from the analysis. This contributes to the degree of

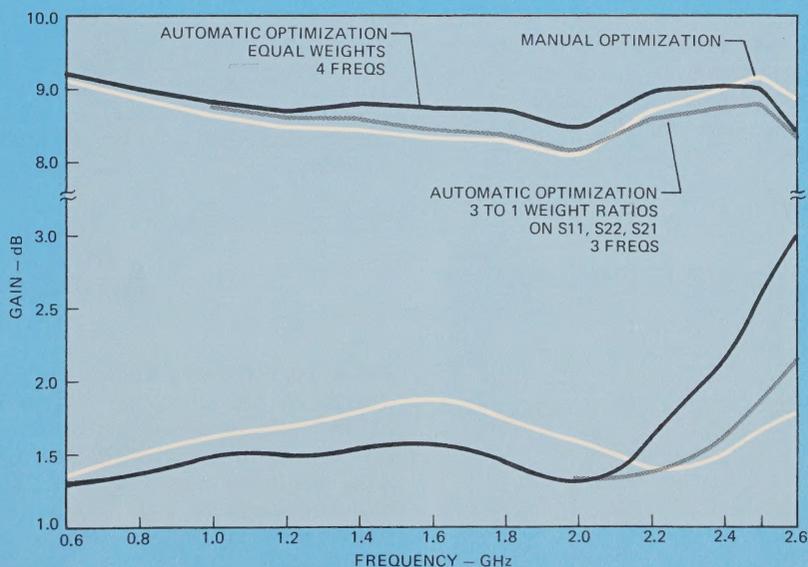


Figure 8. Automatic optimization.

accuracy of the approximation; but, its purpose is to obtain, by simple calculation, what degradation can be expected. This type of approximation is particularly useful when it is desired to determine the influence of changing the feedback elements on noise figure to obtain different gain levels.

When the gain of the bipolar transistor is significantly reduced by feedback, its open loop noise characteristics change such that the amplifier does not respond as readily to changes in source impedance. The transistor will generally operate better when essentially matched to its input impedance. As discussed earlier, the cascaded amplifier must have close to a 50-ohm input and output impedance. For the purpose of the following analysis it will be assumed that the input and output impedance is 50 ohms.

The equivalent circuit for the transistor circuit in Figure 9 can be changed to that shown in Figure 10 for determining the resistive losses caused by the feedback elements. In effect, the emitter feedback resistor,  $R_e$ , can be considered a loss element in series with  $R_1$ , the effective input resistance of the transistor. The emitter resistor can be treated as a voltage-dividing element where the signal-to-noise ratio is increased by this voltage division.

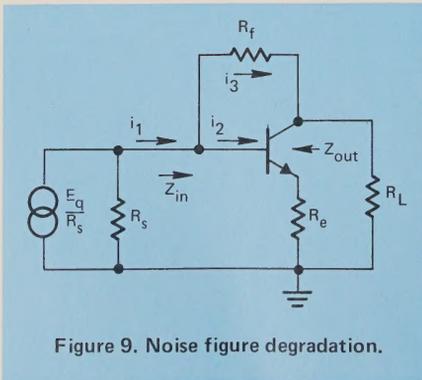


Figure 9. Noise figure degradation.

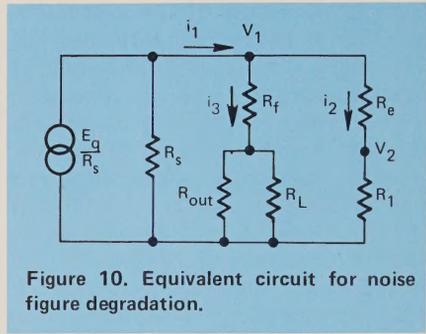


Figure 10. Equivalent circuit for noise figure degradation.

Letting  $Z_{in} = R_{out} = R_L = 50$  ohms and solving for  $R_1$ ,

$$R_1 = \frac{(R_f + 25) 50}{R_f - 25} - R_e \quad (1)$$

The input power to the network when  $Z_{in} = 50$  ohms is  $50i_1^2$ . The power delivered to  $R_1$  is  $R_1(i_2)^2$ . Taking the ratio of delivered power to input power yields:

$$\text{Power ratio} = \frac{R_1 (i_2)^2}{50(i_1)^2} \quad (2)$$

Using current division, and solving for  $i_2$ , gives:

$$i_2 = \frac{i_1(R_f + 25)}{R_f + 25 + R_e + R_1} \quad (3)$$

Substituting equation 3 into equation 2 yields:

Power ratio = 10 log

$$\left[ \frac{(R_f + 25)^2 R_1}{50(R_f + 25 + R_e + R_1)^2} \right] \quad (4)$$

For an amplifier with an  $R_e$  of 12 ohms and  $R_f$  of 240 ohms, the degradation of the noise figure calculates to 1.85 dB. At the other end of the scale a low noise amplifier might have an  $R_f$  of 400 ohms and an  $R_e$  of zero ohms, from which only 0.54-dB degradation is calculated.

Empirical data show that using equations 1 and 4 can be made slightly

more accurate by adding a correction factor of 0 dB for a 20-dB or higher gain stage and 0.5 dB for a 7.5-dB gain stage, with linear interpolation in between.

## Output Power

The output power of the TO-8 feedback amplifier is a direct function of the collector current. The great majority of cascadable amplifiers use direct capacitively coupled outputs, i.e., circuits without impedance transforming devices and, as a result, are voltage limited. Nonsymmetrical clipping occurs on the cutoff portion of the AC swing, as shown in Figure 11. The steep AC load line is a result of

the direct coupling to the 50 ohm load impedance. For greater output power at a given collector current, the equivalent AC load impedance must be raised. This reduces the slope of the AC load line, thereby providing greater AC swing in the cutoff part of the cycle. This effect can be observed by comparing  $V_p$  to  $V_p'$  in Figure 11.

For wideband operation and good impedance matching, it is generally not practical to use a transforming device, except for the VHF and lower UHF frequency ranges. The WJ-PA1, PA2, and A87 use autoformers in the collector circuit for higher and more efficient output power, but are limited to below 500

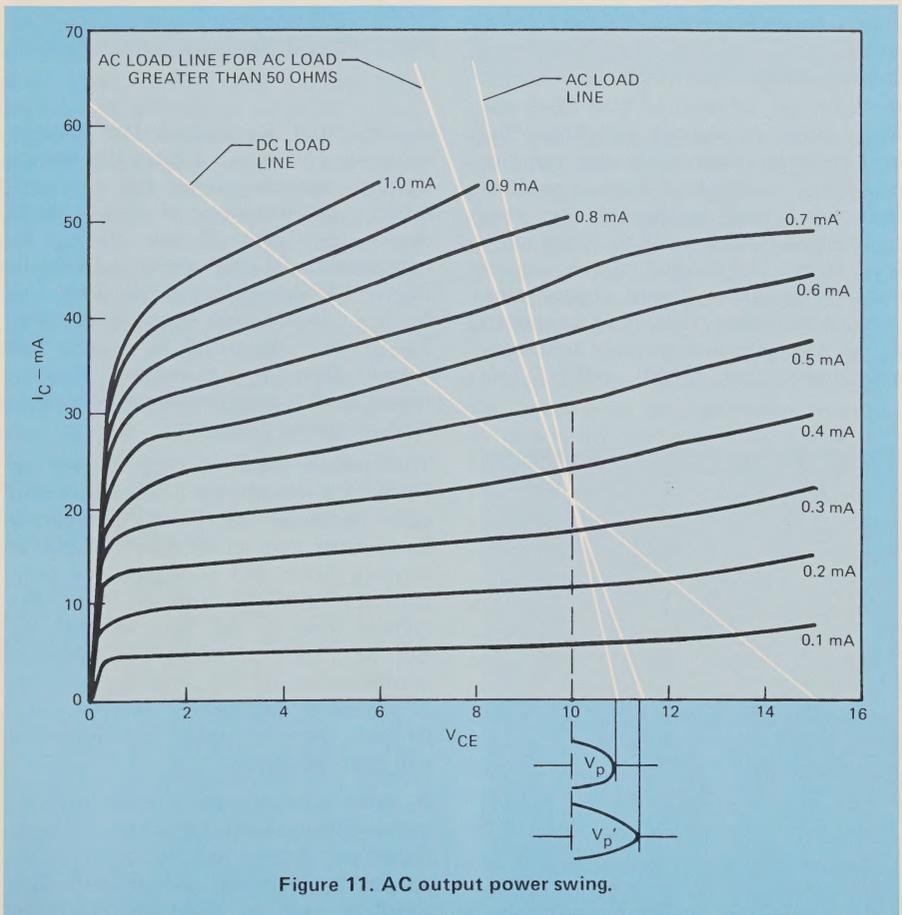


Figure 11. AC output power swing.

MHz operation. Because of the sensitivity of the output power to the AC load line, the voltage limiting is a direct function of the load impedance. Higher than 50 ohms for the load will result in greater than the nominal output power, and lower than 50 ohms will reduce the available output power.

## DC Biasing

Depending on RF requirements and cost considerations, several schemes are used to provide the DC biasing in the cascaded amplifiers. In the low-cost scheme shown in Figure 12, the collector feedback resistor serves the dual purpose of providing base bias and collector-to-base shunt feedback. This approach is used in TO-5 packages and is represented by the W-J EA series, such as the EA2 amplifier. Compromised performance results because the ultra-small size does not allow input or output matching. The unit operates essentially with resistive matching, except that a series peaking inductor is used in the collector line. In addition, the value of  $R_1$  is typically less than 150 ohms, which causes substantial noise figure degradation because the value of  $R_f$  must be chosen to set the gain and provide a reasonable input and output match. This

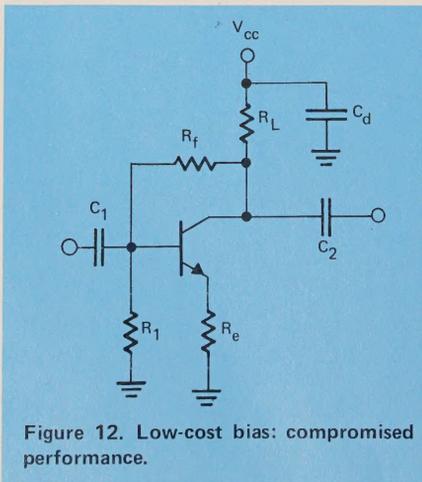


Figure 12. Low-cost bias: compromised performance.

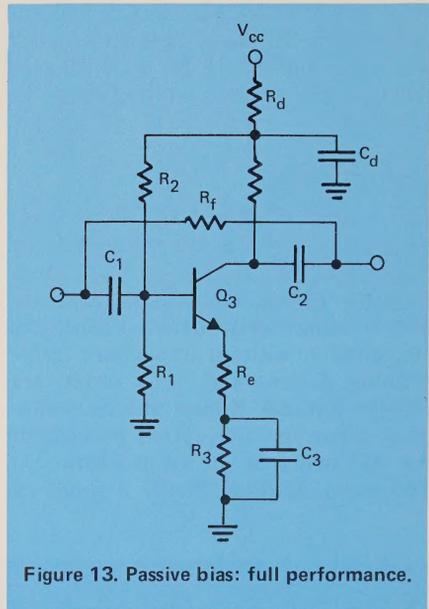


Figure 13. Passive bias: full performance.

requires that  $R_1$  have a low enough value to enable it to pass the excess bleeder current. Since the compromised performance of the circuit shown in Figure 12 has no  $R_3$ , in comparison to the circuit shown in Figure 13, the voltage at the base will be low, which also necessitates that  $R_1$  be low. In order to reduce the current flow,  $V_{CE}$  is set low, thereby reducing the maximum available gain in this type of design.

The passive bias circuit shown in Figure 13 overcomes noise figure and gain problems by using a separate  $R_f$  resistor and an emitter resistor to provide both DC feedback for temperature stability and to raise the voltage level at the base so that  $R_1$  can be larger in value. The parallel combination of  $R_1$  and  $R_2$  provide a minimum amount of shunting and its accompanying noise figure increase and gain reduction.

A third scheme uses a PNP low frequency transistor to bias the RF transistor, as shown in Figure 14. This approach virtually eliminates base shunting and, in addition, does not

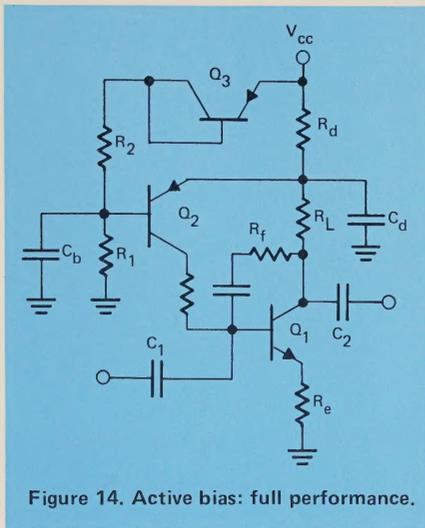


Figure 14. Active bias: full performance.

have an element in the emitter circuit which minimizes degeneration because of additional parasitic inductance in the emitter leg. This circuit is temperature-stable because if  $I_c$  in the RF transistor goes up, the voltage across the sense resistor,  $R_d$ , will go up, causing  $V_{BE}$  across  $Q_2$  to go down. This action reduces the base drive to  $Q_1$ , causing  $I_c$  to go down. When  $R_d$  is less than about 300 ohms, it is usually necessary to include  $Q_3$  to track the base-emitter voltage of  $Q_2$  over temperature as the differential voltages on  $R_d$  can become comparable in magnitude to the normal base-emitter voltage change that occurs as a function of temperature.

The circuit of Figure 15 is used for the higher-level cascadable amplifiers, which differ from the type shown in Figure 14 by the addition of an RF choke. The choke does not limit current, and for higher-level units, particularly if it is not desired to raise  $V_{cc}$  above 15 volts, it does not shunt the output and absorb RF power. Most W-J amplifiers that provide +13 dBm or greater output power utilize an RF choke for power supply decoupling.

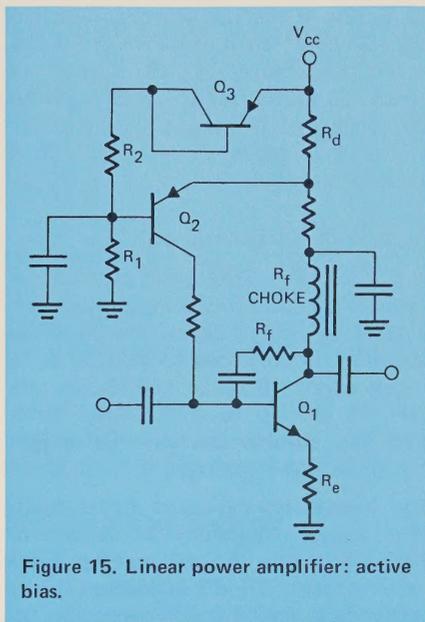


Figure 15. Linear power amplifier: active bias.

One method of providing bias for two cascaded RF stages in a single TO-8 is demonstrated in Figure 16. Here, the active bias circuit of Figure 14 is expanded to provide bias from a single transistor for both RF transistors. The limitation of this circuit is that the collector current of each RF transistor ( $Q_1$  and  $Q_2$ ) must be close enough in value so that each transistor will receive its proportionate amount of correcting currents from  $Q_3$  when the circuit is operated over wide temperature ranges.

To summarize these bias schemes, the circuit in Figure 12 is used where minimum cost and size are the dominant factors. The circuit in Figure 13 is used where higher performance is desired, but the frequency of operation is usually limited to 500 MHz or below. The circuit will be less sensitive to sub-band oscillators (see section on sub-band oscillations), since no active device is used in the bias circuit.

The bias circuit shown in Figure 14 is not frequency-limited, and offers the maximum gain obtainable from the RF transistor. The circuit of Figure

15 is used where it is desired to obtain higher power than possible from the circuit of Figure 14. The circuit of Figure 16 minimizes the area required for high performance applications when two stages in a single TO-8 package are used.

### Available Performance

A plot of the highest gain available in a single TO-8 as a function of frequency is given in Figure 17. Over 30 dB is currently available up to 500 MHz, in excess of 20 dB up to 1500 MHz, and over 10 dB up to 3200 MHz. Each amplifier shown uses two RF stages in a single TO-8 package.

The lowest typical noise figure available versus frequency is shown in Figure 18, with 2.5 dB maximum up to 500 MHz, 3.0 dB maximum up to 1000 MHz, and 4.0 dB maximum up to 2000 MHz.

The highest output power available versus frequency is shown in Figure 19, with typically greater than +26 dBm up to 100 MHz, greater than +24 dBm to 1200 MHz, and greater than +20 dBm to 2000 MHz.

### Stability

Each W-J cascaded amplifier has been designed to be unconditionally stable with load VSWR's of any phase angle. In order to realize inherent stability, certain installation procedures and methods must be employed.

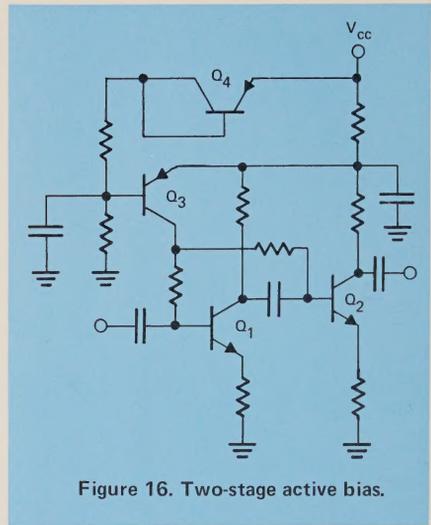


Figure 16. Two-stage active bias.

Instability falls typically into two types: 1) Sub-band oscillations in the 100 kHz to 1 MHz region and, 2) out-of-band oscillations above the frequency band. Occasionally in-band oscillations can occur, but are much more rare and are usually related to poor circuit board grounding.

### Grounding

The necessity for good grounding cannot be over emphasized. The following measures provide acceptable grounding. First, the entire TO-8 case should be completely grounded across its entire surface to the circuit board. The circuit board,

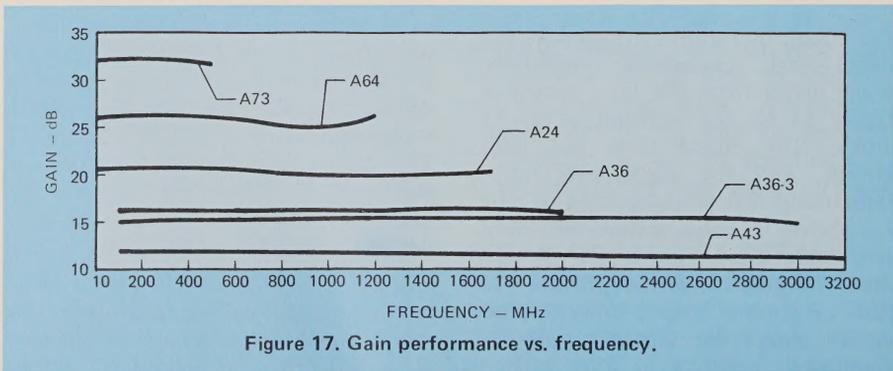


Figure 17. Gain performance vs. frequency.

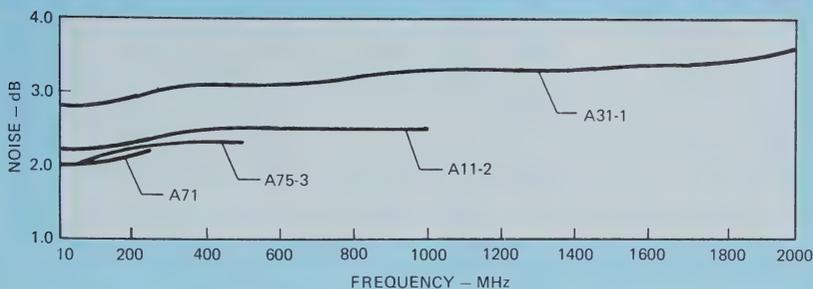


Figure 18. Noise performance vs. frequency output.

in turn, must be properly grounded on both sides. Shown in Figure 20 are both sides of a typical TO-8 circuit board used for mounting the TO-8's. The board ground planes are connected on the edges by the mounting screws and the case to which it is mounted. In addition, a series of small plated through-holes are placed parallel to the RF circuit trace for the purpose of reducing the inductive reactance between the two ground planes. Insufficient ground connections can result in midband or high-end out-of-band spurious responses. The W-J TO-8 mounting kit (No. 251050-001) is the preferred technique for holding the TO-8 to the ground plane. A conductive epoxy ring that fits around the periphery of the bottom of the

case is also an acceptable method, where space is at a premium.

### Sub-band Oscillations

Sub-band oscillations are most frequently related to insufficient grounding of the RF signal, which is generally solved by using a  $0.1 \mu\text{f}$  capacitor to ground on the  $V_{CC}$  port. Since each W-J TO-8 amplifier is internally decoupled with a minimum value of  $0.01 \mu\text{f}$ , adding only  $.01 \mu\text{f}$  to the  $V_{CC}$  port is usually not sufficient to totally suppress sub-band oscillation.

Shown in Figure 21a is the noise floor of two cascaded A74 amplifiers which have a total gain of approximately 60 dB. The horizontal scale is 0.5 MHz per division. The noise spike appearing at approximately 400 kHz reveals a

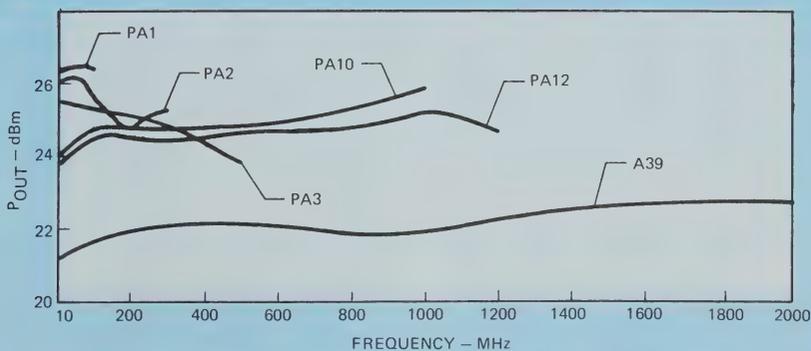


Figure 19. Output power performance vs. frequency.

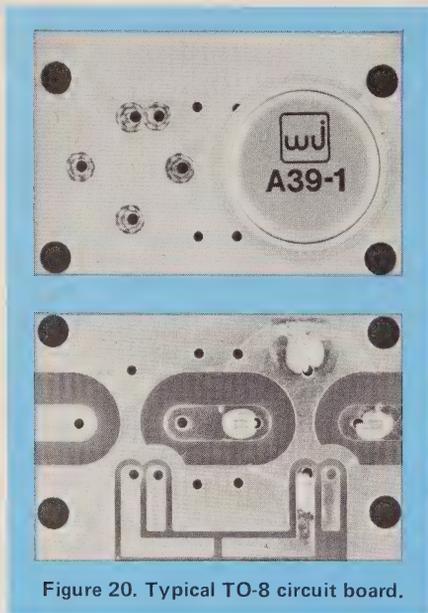


Figure 20. Typical TO-8 circuit board.

tendency for instability. The decoupling capacitor is  $0.05 \mu\text{f}$ . The results for a  $1 \mu\text{f}$  decoupling capacitor are shown in Figure 21b. Note the substantially reduced noise level at 400 kHz.

If the bias line from the power supply is not properly shielded sub-band oscillations can occur, particularly when high gain levels of 60 dB and higher are used. In Figure 21c,  $0.05 \mu\text{f}$  of decoupling and five inches of unshielded power supply line is used. Figure 1 should be used for comparison since all conditions were the same except for the shielding. Figure 21a and 21b used shielded power supply lines. Note the "comb" generator effect that is created by lack of shielding.

If the cascaded or single unit has a potential for oscillation, as is shown in Figure 21a, under terminated conditions, certain load conditions may cause sub-band instability, such as open or short circuit input conditions. The higher the cascaded gain the higher the potential for oscillation. In this case, it is particularly important

that sufficient decoupling is used. Figure 21d shows what can happen at 60 dB of gain when the input is open-circuited and when there is  $0.05 \mu\text{f}$  on the bias line. The addition of  $1 \mu\text{f}$  to the  $V_{\text{CC}}$  port eliminates the oscillation under the open circuit load condition, as shown in Figure 21e.

## Radiative Feedback Oscillation

Radiative feedback oscillation can occur when a high gain chain is used where the RF output voltage is in phase with the input voltage (even number of stages), and the output signal radiates back to the input because of the antenna action of both the output and input circuit traces. The solution here is simply to keep output and input traces as short as possible and as far from each other as possible. For gain of over 60 dB, it is advisable to shield the input from the output by using a separate cavity for the input gain stages.

## Out-of-band High Frequency Oscillators

High frequency oscillations can occur as a result of poor circuit board and amplifier case grounding. In addition, if the grounding is marginal, the units may be more susceptible to oscillation under certain load conditions. A bare lead of only one-half inch in length used to connect the RF signal from its source to the input port of the amplifier may result in oscillations. It is much better to use a properly grounded circuit board with a microstrip transmission line or coax cable. Any open lines, whether on the  $V_{\text{CC}}$  port or in the RF signal path may require ferrite beads to prevent them from acting as radiators or receptors.

## High-Performance Application

An example of a high performance application for the TO-8's is shown in Figure 22. This is a mixer pre-amplifier assembly referred to as a converter. It uses standard unselected TO-8 amplifiers in conjunction with

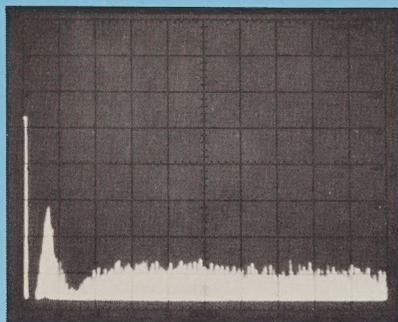


Figure 21A. Noise floor vs. frequency; 0.05  $\mu$ fd decoupling.

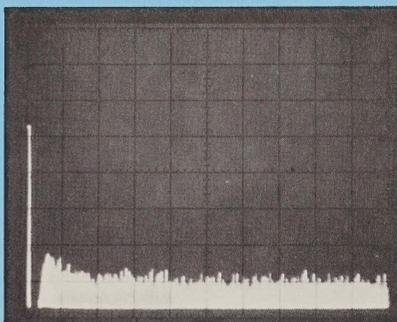


Figure 21B. Noise floor vs. frequency; 1.0  $\mu$ fd decoupling.

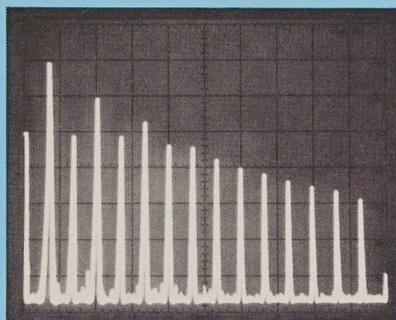


Figure 21C. Unshielded power supply bias lines.

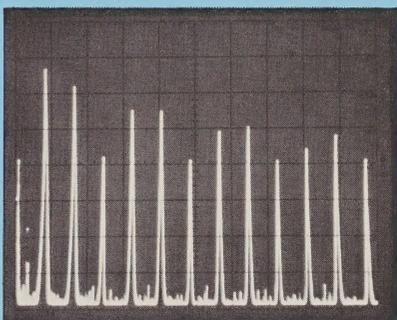


Figure 21D. Open circuit input; 0.05  $\mu$ fd decoupling.

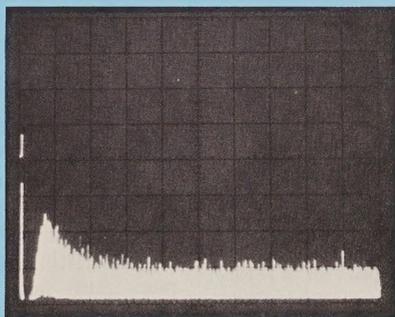


Figure 21E. Open circuit input; 1.0  $\mu$ fd decoupling.

several other processing components, including a WJ-M15 mixer. One of its key features is unit-to-unit phase match that is typically within  $\pm 3$  degrees. This tight phase control is possible in part because of the consistency in phase offset provided by

the TO-8 cascable amplifier. The physically small size of these units reduces the degree of phase shift variations in each unit. Phase adjustment is provided by adjusting a delay line in both the IF port and the LO port of the mixer. In addition, the

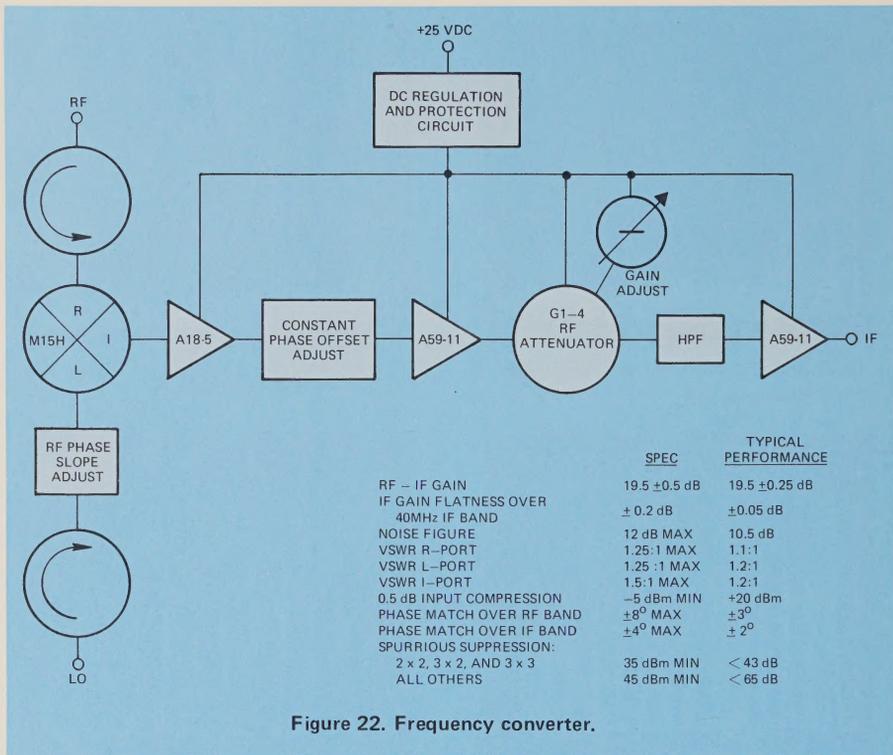


Figure 22. Frequency converter.

M15 mixer, because of its uniconstruction, offers excellent phase shift consistency. The adjustment in the LO port is more sensitive to phase shift than the IF port by the ratio of LO frequency to IF frequency, hence, minor changes in mixer variations can have a strong impact on the phase offset through the unit.

The frequency converter, shown in Figure 22, is used as a part of a large radar system that combines computers, radars, and missiles to provide a complete defensive shield for ships at sea. The radar system can see in all directions simultaneously and can automatically detect, track, and respond with large numbers of aircraft and missiles while maintaining continuous surveillance.

### Conclusion

The TO-8 cascable amplifier has allowed the RF system designer to

concentrate on the intricacies of the system design, with a maximum degree of flexibility. The ultrabroad bandwidth of the units allows a much smaller number of models to work in a large number of applications. The building block principle also allows the flexibility needed for the many different types of requirements. The TO-8 cascable amplifier's small size also contributes directly to its usefulness and has made possible much higher density packaging, which opens the door for new applications, such as smart missiles or jeep-carried radar systems that previously were impossible because of the large bulk needed without TO-8 type amplifiers.

### Acknowledgement

The author wishes to thank Jay Gaertner for supplying the data for many of the curves and spectrum analyzer results.

**Author: Daniel L. Cheadle**



Daniel L. Cheadle is currently Head of the Mixer and Amplifier Section within the Component Engineering Department. In this capacity, he is responsible for designing and directing the design and development of cascadable amplifier products, and the direction of the production project engineering responsibility for the product line. In addition, he is responsible for the direction of microwave mixer design and development.

Prior to this position he was project engineer for the TO-8 cascadable amplifier product line and has been responsible for the design of numerous models that cover frequencies up to 3200 MHz, power levels up to +26 dBm, gain levels up to 33 dB and noise figures below 2.5 dB. He has been the chief RF circuit design engineer for the product line from the beginning of its development within Watkins-Johnson Company.

Before his work in amplifiers he did product development on lumped element mixers, with a specialty in the ultra-high level regime. For example, he was directly responsible for the design and development of the M9E/M9D mixers, and later designed the TO-8 thin-film M9G/M9H and M2E high-level mixers. He has published several articles related to mixers, with emphasis on distortion-related topics.

Mr. Cheadle received a BSEE and MSEE from San Jose State University.

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## Manufacturing and Office Locations

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#### SALES OFFICES

##### CALIFORNIA

Watkins-Johnson  
3333 Hillview Avenue  
Palo Alto 94304  
Telephone: (415) 493-4141

Watkins-Johnson  
2525 North First Street  
San Jose, 95131  
Telephone: (408) 262-1411

Watkins-Johnson  
831 South Douglas Street  
Suite 131  
El Segundo 90245  
Telephone: (213) 640-1980

##### DISTRICT OF COLUMBIA

Watkins-Johnson  
700 Quince Orchard Road  
Gaithersburg, Md. 20760  
Telephone: (301) 948-7550

##### FLORIDA

Watkins-Johnson  
325 Whooping Loop  
Altamonte Springs 32701  
Telephone (305) 834-8840

##### MARYLAND

Watkins-Johnson  
700 Quince Orchard Road  
Gaithersburg 20760  
Telephone: (301) 948-7550

##### MASSACHUSETTS

Watkins-Johnson  
5 Militia Drive  
Suite 11  
Lexington 02173  
Telephone: (617) 861-1580

##### PENNSYLVANIA

Watkins-Johnson  
385 Lancaster Avenue  
Haverford 19041  
Telephone: (215) 896-5854

##### OHIO

Watkins-Johnson  
2500 National Road  
Suite 200  
Fairborn 45324  
Telephone: (513) 426-8303

##### TEXAS

Watkins-Johnson  
9216 Markville Drive  
Dallas 75243  
Telephone: (214) 234-5396

### International

##### ITALY

Watkins-Johnson Italiana  
S.p.A.  
Piazza G. Marconi, 25  
00144 Roma-EUR  
Telephone: 59 45 54  
Telex: 63278  
Cable: WJROM-ROMA

##### GERMANY, FEDERAL REPUBLIC OF

Watkins-Johnson  
Manzingerweg 7  
8000 Muenchen 60  
Telephone: (089) 836011  
Telex: 529401  
Cable: WJDBM-MUENCHEN

##### UNITED KINGDOM

Watkins-Johnson  
Shirley Avenue  
Windsor, Berkshire SL4 5JU  
Telephone: Windsor 69241  
Telex: 847578  
Cable: WJUKW-WINDSOR

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