high dynamic range receiver input stages

A new look at the input stages of high-frequency communications receivers including some circuits for improving the IMD performance over that of any present amateur equipment

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Various efforts have been made in the past to build high-frequency communications receivers with wide dynamic range to combat the effects of cross-modulation and blocking. With the high level of galactic and man-made noise which is present up to about 30 MHz, receiver designers generally agree that a receiver with a 10 dB noise figure is more than adequate for high-frequency communications. This is equivalent to a sensitivity of 0.2 μV for 10 dB signal-plus-noise to noise (50-ohm input impedance, 2-kHz bandwidth). In most cases, on the high-frequency bands, the man-made noise which is picked up by the antenna is greater than 0.07 μV, so the extraordinary sensitivity of a receiver with a noise figure lower than 10 dB is no great advantage. While this technique was initiated ten years ago by my company, Rohde & Schwarz, in Munich, amateur equipment designers have only recently started thinking along the same lines.

A noise figure of slightly less or equal to 10 dB can be readily achieved by using one of the new hot-carrier diode double-balanced mixers, such as the Minilabs SRA3H or SRA1H, which is followed by a low-noise i-f amplifier ahead of a crystal filter. In practice this means that no rf amplifier stage is required. In vacuum-tube receivers one of the primary applications for the rf input stage was to provide agc but this can now be obtained (with negligible losses) with a PIN diode attenuator.[1,2]

The first amateur equipment to use this new design technique is the Atlas 180 and its later models. There are some indications that this circuit is derived from the SouthCom SC130 Man-Pack Transceiver* and the AN/PRC-104 of Hughes Aircraft, which both use the same technique. The main disadvantage of this rf input circuit is that it is practically impossible to suppress feedthrough energy from the oscillator to the antenna to 15 μV or less, a requirement of European regulatory agencies.

Some military and systems-oriented receivers still require noise figures of 6 dB or less because they are used with inefficient antennas (whip antennas with loading coils for marine, mobile or portable use, for example). Since it is difficult for the commercial communications equipment manufacturer to foresee whether the receiver will be used with a good antenna or a poor one, coupled with the requirement for low oscillator feedthrough, in many cases rf amplifier stages are still required.

Recently the German amateur journal, *cq-DL*, published an extensive test report on the Atlas 180[3] which showed fairly poor dynamic range with respect to what would be expected from a receiver with a double-balanced mixer input and high-power i-f stage before the crystal filter. As pointed out in a previous article,[4] to obtain the specified performance every mixer must see a purely resistive load at *both* the i-f and image frequency. An analysis of the SouthCom and Atlas circuit reveals a tuned circuit between the double-balanced mixer and the i-f stage. Since the mixer is not properly terminated at the image frequencies, this results in a loss of at least 15 dBm for the mixer's third-order intercept point, enhancing intermodulation distortion products.

table 1. Third-order intercept point of Minilabs SRA1H high-level double-balanced mixer with various terminations.

termination	intercept point
50-ohm resistance	30 dBm
Narrow-band resonant circuit	8 dBm
Heavily damped resonant circuit	17 dBm
Elliptical bandpass circuit	21 dBm
Amplifier (Z_i = 50 ohms ±10°, 1-80 MHz)	23 dBm
Power fet (Z_i = 50 ohms ±5°, 1-108 MHz)	30 dBm

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mixer circuits

A recent survey published in Germany[5] showed the results of a series of tests regarding the creation of unwanted spurious in double-balanced mixers as a function of termination. All tests were run with a high-level Minilabs SRA1H mixer which requires a local-oscillator input of 23 dBm (200 milliwatts). The results of these tests are listed in **table 1**.

Although the Minilabs SRA1H is capable of providing a third-order intercept point of 30 dBm, **table 1** indicates that you can lose as much as 22 dB of dynamic range because the mixer is not properly terminated. Therefore, to obtain maximum performance, it is essential to build an input stage which has the required input impedance from dc up to more than 100 MHz. This stage must also have an intercept point of 25 dBm (the double-balanced mixer has about 5 dB loss so the mixer output is 25 dBm at 30 dBm input).

Extensive tests with various crystal filters have been conducted in the past and it was found that the crystal filter's input and output must be terminated beyond its normal frequency range of operation. This is because crystal filters, such as the KVG XF9B, at frequencies not too far from the center frequency, exhibit impedances between 5 and 10 kilohms (see **fig. 1**). These resonant effects would significantly reduce the dynamic range of the preamplifier if the filter was not properly terminated.

The best solution to the problem of a high dynamic range preamplifier is a high current field-effect transistor, type CP643 (Teladyne Crystalonics), in the circuit of **fig. 2** where the input resistors can be adjusted to set the dc bias so the input impedance is exactly 50 ohms.[5] This circuit has been designed for optimum dynamic range regardless of oscillator feedthrough and does not show any input selectivity which must, of course, be added. In most cases input selectivity can be provided by a 1.6-MHz highpass filter and a 31-MHz lowpass filter, so the losses there do not have to be considered.

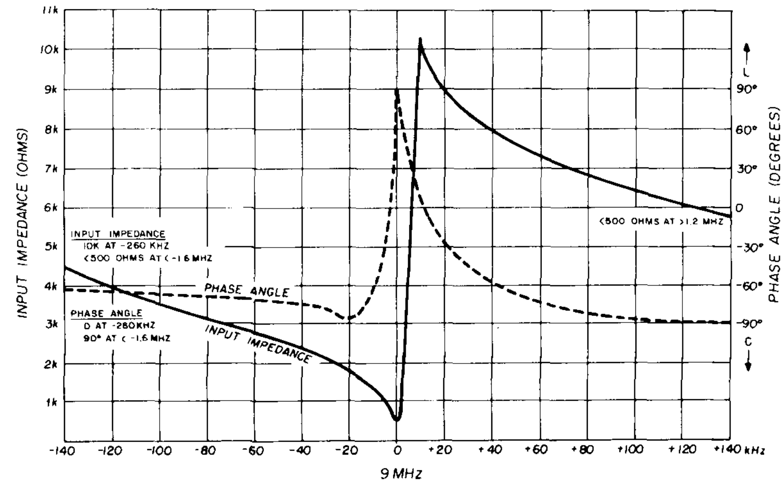


fig. 1. Input impedance and phase angle of the 9-MHz KVG XF9B crystal filter when terminated in 560 ohms in parallel with 33 pF.

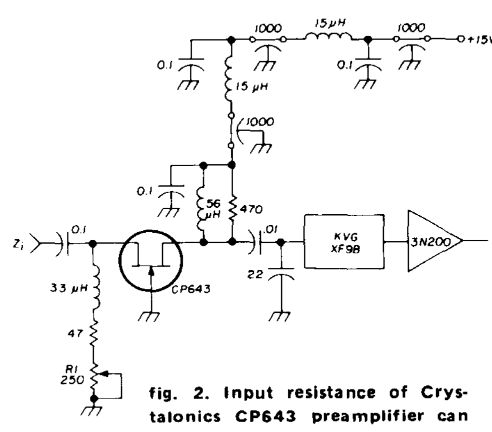


fig. 2. Input resistance of Crystalonics CP643 preamplifier can be adjusted to 50 ohms by proper setting of the 250-ohm potentiometer, R1. Current drain is 30 mA.

The complete rf input circuit, showing the CP643 high dynamic range preamplifier, double-balanced mixer, 2N5109 oscillator injection amplifier and 3N200 i-f amplifier is shown in **fig. 3**. The injection amplifier accepts about 200 mV input voltage from the local oscillator and amplifies it to the required level for the double-balanced mixer. The grounded-gate CP643 has 14 dB gain and is matched to the input of the KVG XF9B crystal filter. The 3N200 i-f amplifier has enough gain and agc action for most receiver designs.

push-pull rf amplifier with wide dynamic range

In many wideband, high dynamic range applications such as antenna distribution amplifiers, input rf amplifiers are required which combine extremely low distortion with low noise figure. In the past both voltage and current feedback have been used to counteract voltage and current distortion. The disadvantage of these circuits is that a stable input impedance can be achieved only over a relatively narrow bandwidth.[6]

Recent research has resulted in a new

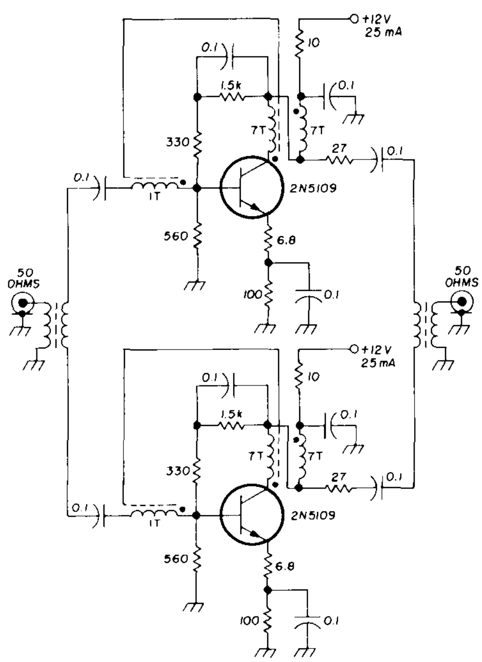


fig. 4. Vhf power transistors are used in push-pull circuit to obtain wide dynamic range shown in fig. 5. Transformers are trifilar wound on Indiana General F625-9-TC9 toroid cores.

wideband amplifier design[2,7] which has extremely low vswr at both input and output as well as low noise figure. The

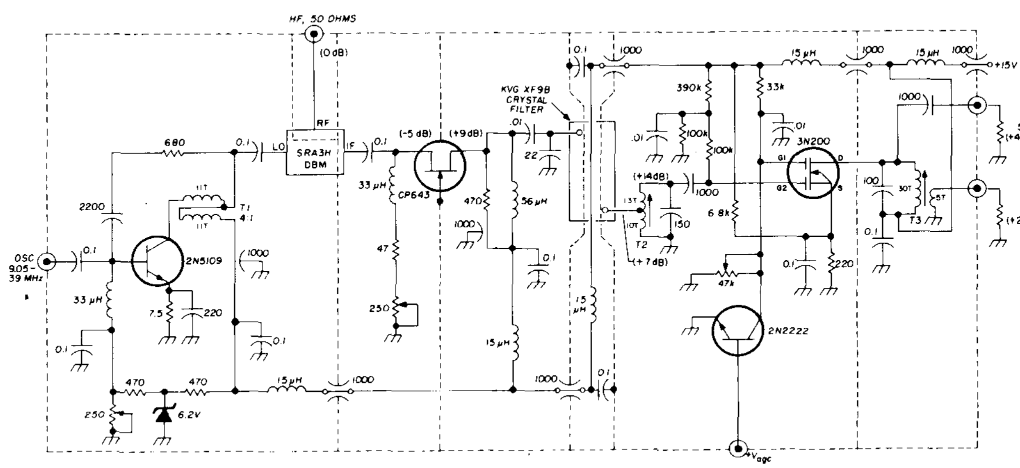


fig. 3. Double-balanced mixer circuit which has a third-order intercept point at +30 dBm. Oscillator requirement is −1 to +2 dBm (200 to 280 mV across 50 ohms). Agc range is greater than 50 dB. Levels shown in parenthesis are for zero dBm (224 mV) at input and zero agc voltage.

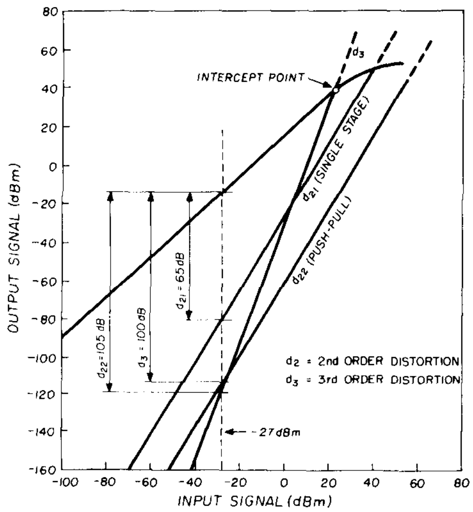


fig. 5. Performance of the push-pull rf amplifier shown in fig. 4. With an input of −27 dBm (two-tone signal, 20 mV each), gain is 12 dB, third-order distortion products are down 100 dB and second-order IMD is down 105 dB. Third-order intercept point occurs at an input of about 22 dBm.

second-order intermodulation products can be suppressed nearly 40 dB (over a single stage) by the push-pull arrangement shown in **fig. 4**. Two linear vhf power transistors are used in the circuit, and depending upon the large-signal handling requirements, either the 2N5109 (RCA) or BFR95 (Amperex) may be used. Both of these devices have an F_T of 1600 MHz.

This circuit provides about 11 dB gain and exhibits exceptional freedom from second- and third-order intermodulation products, as plotted in **fig. 5**. The third-order intercept point occurs at an input of about 22 dBm. Three types of feedback are used: *current* feedback through the unbypassed 6.8-ohm emitter resistor, *voltage* feedback through the unbypassed 330-ohm base-to-collector resistor, and *transformer* feedback through a third winding on the wideband transformer to stabilize the input and output impedance. A mathematical analysis of this circuit is presented for interested readers in the appendix.

table 2. Measured third-order intercept point of several commercial high-frequency receivers.

receiver	intercept point
Yaesu FT101	−21.5 dBm
Ten-Tec Argonaut	−19.5 dBm
Collins KWM2/S-line	−10.0 dBm
Signal 1 CX7	−5.0 dBm
Collins R390A	−4.5 dBm
Atlas 180/210	3.0 dBm
Collins 65S1	13.0 dBm
Racal RA1772	28.0 dBm
Martin rf front end (fig. 3)	30.0 dBm

summary

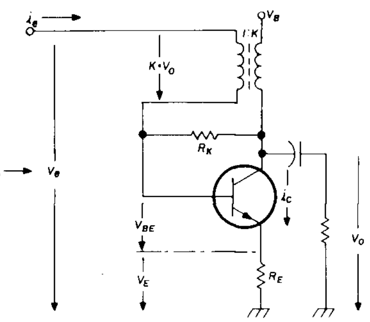
With very little effort the third-order intercept point of high-frequency receiver input stages can be increased far beyond the values of any commercial equipment now on the market. **Table 2** shows the third-order intercept point of several popular receivers presently being used by amateurs.

references

1. Ulrich L. Rohde, "Eight Ways to Better Receiver Design," *Electronics*, February 20, 1975, page 7.
2. Ulrich L. Rohde, "Zur Optimalen Dimensionierung von Kurzwellen-Eingangsteilen," *Internationale Elektronische Rundschau*, November, 1973, page 244.
3. Thomas Moliere, "Der Transceiver Atlas 180 - Testbericht," *cq-DL*, March, 1975, page 130.
4. Peter Will, "Reactive Loads — The Big Mixer Menace," *Microwaves*, April, 1971, page 38.
5. M. Martin, "Extrem lineares Empfaenger-Eingangsmodul mit grossem Dynamikbereich und sehr geringen Intermodulationsverzerrungen," *Internationale Elektronische Rundschau*, April, 1975, page 73.
6. A. Hoenike, "Dimensionierungsfragen beim Entwurf klirrarmer Transistorverstaerker," *Nachrichtentechn. Z.*, May, 1966, page 287.
7. K.H. Eichel, "Einfache Methode zur Erzielung eines konstanten Eingangswiderstandes bei Breitbandverstaerkern," *Internationale Elektronische Rundschau*, February, 1973, page 45.

appendix

Since voltage and current feedback may result in input and output impedances which may not suit the design requirements, a wideband toroidal transformer with a high-permeability core (such as Indiana General, F625-9-TC9) can be used to arbitrarily set these impedances. The schematic below



shows an amplifier using a transformer with voltage and current feedback. The input voltage, v_e, input current, i_e, and input impedance, Z_e, are given by the following equations

$$v_e = (k \cdot v_o) + v_{be} + v_e$$

$$i_e = (v_{be} \cdot Y_{11}) + v_{cb}/R_k$$

$$Z_e = \frac{(k \cdot v_o) + v_{be} + v_e}{(v_{be} \cdot Y_{11}) + v_{cb}/R_k}$$

So long as the operating frequency is well below f_T, emitter and collector current are the same. Therefore, the input impedance of the stage will be

$$Z_e = R_k \left(\frac{k + A}{1 - A}\right)$$

where $$A = \frac{R_E\,(R_k + R_L)}{(R_E + R_k)\,R_L}$$

and the required value for the voltage feedback resistor, R_k, is given by

$$R_k = Z_e \left(\frac{1 - A}{k + A}\right)$$

the amplification of the stage is given by

$$A = \frac{1}{k + \dfrac{R_E\,(1 + R_k/R_L)}{R_k + R_E}}$$

Since it is advantageous to have a 50-ohm input impedance, and the output impedance of the stage is approximately 150 ohms, the collector winding is split to build a 4:1 transformer. Under these conditions the input impedance, Z_e, output impedance, Z_o, and voltage gain, A, are given by

$$Z_e \approx (k \cdot R_k) + \frac{R_E}{2}$$

$$Z_o \approx \frac{R_E}{2k}$$

$$A = \frac{1}{k\left(1 + \dfrac{R_E}{2k \cdot Z_L}\right)}$$

The constant k, which determines the turns ratio between the base and collector coil (1:7:7 in **fig. 5**), can only be an integer. To obtain optimum performance in many cases, therefore, one of the values may have to be a compromise. In the circuit of **fig. 5** the output impedance was only 23 ohms so a 27 ohm resistor had to be placed in series to obtain the desired 50-ohm output impedance.

ham radio

capacitance meter

Dear HR:

I have receiver a number of queries regarding the programmable unijunction transistor used in the capacitance meter described in the April, 1975, issue of *ham radio*. The full part number of this device, which is manufactured by Texas Instruments, is A7T6028 (because of space limitations, the package is labeled AT6028). The 2N6027, 2N6028 and 2N6118 are similar.

Although I have not tried the Motorola HEP S9001 programmable unijunction in the circuit, I have letters from K6MYA and WA3IFQ who say they had excellent results with this device. Another possibility is the package of programmable unijunctions available from Radio Shack (part number 276-119).

The two series-connected 0.005 μF capacitors in the circuit may be replaced by a single 0.0025 μF capacitor. This apparently has caused some confusion.

Courtney Hall, WA5SNZ
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