converting a vacuum-tube receiver to solid state

You can reduce drift in older receivers by an order of magnitude– a modification scheme for the BC348 using MOSFET's and JFET's

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The BC-348 receiver is well known for its excellent selectivity and mechanical stability. At the same time, this old iron box can be an instrument of torture for the user. Some of the earlier BC-348's were bedeviled with the "scratchy dial" syndrome. This is a phenomenon of unknown origin and is virtually impossible to cure. It appears when the receiver is tuned and sounds like a thousand demons rubbing a tin washboard. If you're unfortunate enough to have a BC-348 with this ailment, and the receiver has a drift problem, it can be a real chore indeed to keep a signal in the i-f passband.

In resuming an interrupted conversion of the BC-348 begun a dozen years ago, I decided the only way to reduce frequency drift was to reduce internal temperature rise. There are several ways to do this. Reducing B plus takes off seven watts; a transistorized audio circuit eliminates another nine. To reduce drift by ten times, however, the power consumption must be cut by 90 percent. Answer: **all** transistors.

If the conversion includes a class-B audio amplifier, drain under normal conditions would be well under ten watts, and two watts of this would be in the pilot lights. Fine. Now the question is, how to do this without sacrificing any of the features this receiver is noted for?

choosing replacements

The BC-348, -312, and -342 receivers use rf coils with low-impedance plate windings (or taps) and high-impedance grid windings. Bipolar transistors might be used by turning things around (interchanging grid and plate), or junction FET triodes might be used in a cascode circuit, as in the H. H. Scott fm tuners—two JFET's replace one 6K7.

However, the RCA dual-gate MOSFET's, recently put on the market at a reasonable price, have the correct input and output impedances for this use and afford a gain control with lower distortion products. It turns out that the 3N140 MOSFET, at rated current, has about five times the transconductance of the 6K7. The relatively low output impedance of the 3N140 (**table 1**) is not objectionable when used with low-impedance inputs.

Because of the interstage coupling, stage gain in the BC-348 receivers increases from the low to the high end of each band. In some receiver models a gain control potentiometer was ganged to the tuning capacitor. I felt that this was an added complication not needed for amateur work, so I cut the wires going to the pot and those from the pot to the bandswitch inside the coil box.

Tubes I replaced with FET's were the 6K7 rf amplifiers, a 6J7 mixer, and 6C5 local oscillator. Characteristics of these tubes at 18 MHz are given in **table 1** with those of some replacement semiconductors. Problems occurred with rf gain, rather than with interelectrode capacities, because of signals feeding back along the tuning capacitor rotor (from gang 3 to the antenna section, for instance) before other effects showed up. Also, the 3N140 will oscillate enthusiastically at any frequency up to 300 MHz as I found out.

The extra resistors and capacitors in the circuit are mainly to kill various sorts of vhf parasitics. In order to get the rf gain down where it belonged (high rf gain will also cause difficulty with mixer overloading) it was necessary to put capacity voltage dividers

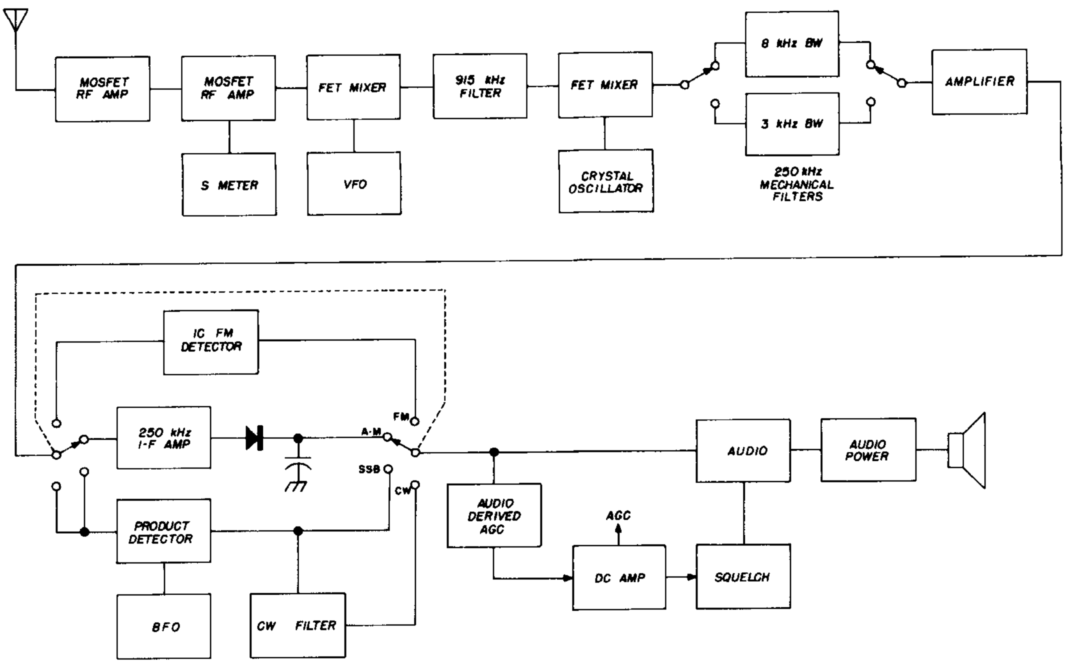


fig. 1. Simplified block diagram of the solid state BC-348 receiver. In the actual receiver, the audio-derived agc circuit is switched out of the system in the a-m mode.

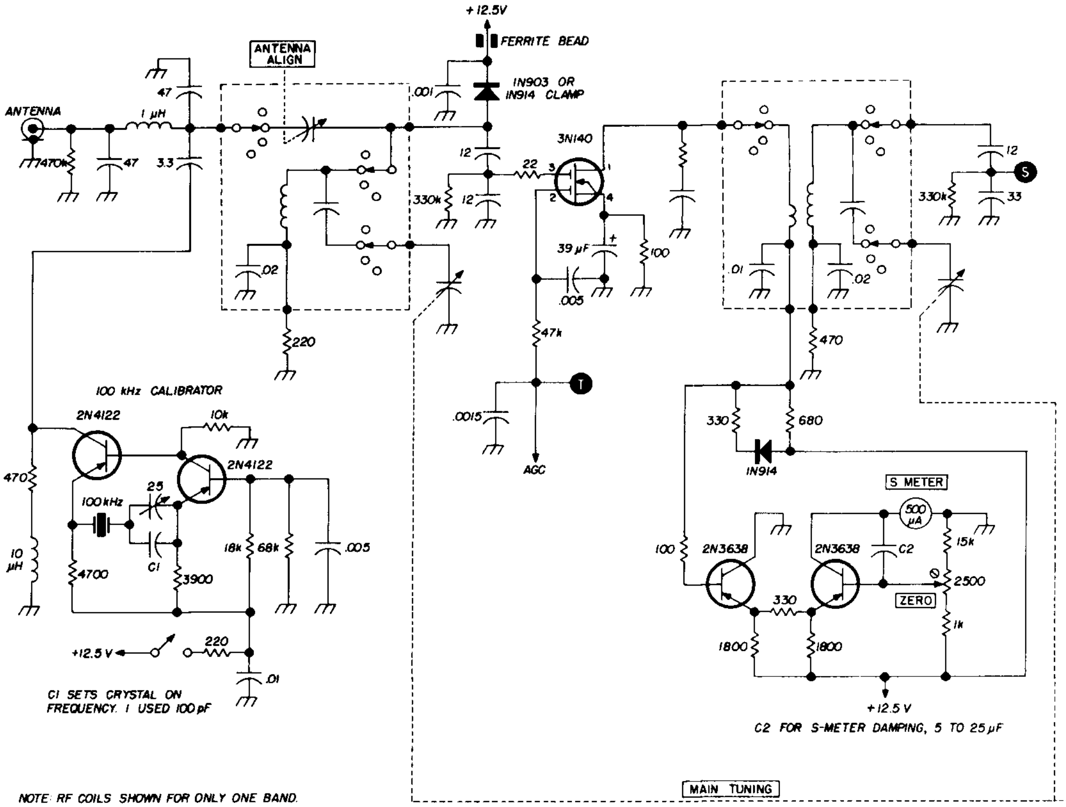


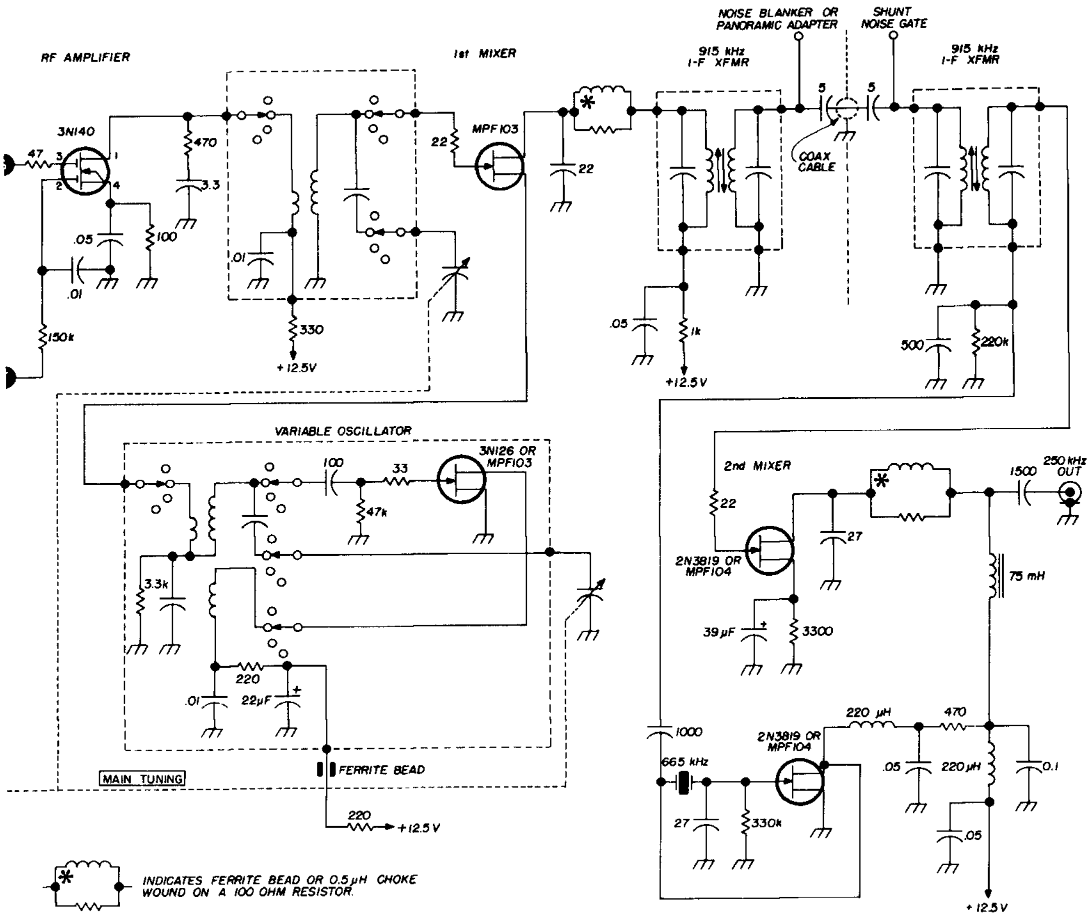
fig. 2. Schematic diagram of the solid-state BC-348. Coils in the rf section are shown for one band only; bandswitch shaft end should be grounded with a spring to remove 144-MHz birdies. Resistors are 1/4-watt composition unless otherwise marked; 1% resistors are 1/8-watt metal-film types (RN55 or RN60). Polarized capacitors less than 25-μF are tantalum types (150D or CS13); aluminum electrolytics are ok if not used for agc or audio coupling.

at the input of each 3N140. Because the semiconductor triode noise is lower than that of the 6K7, the measured noise figure of the modified receiver was somewhat better than the original, even though there wasn't always a peak in noise as the antenna trimmer was turned.

Gain control is provided by changing the bias on the second gate. Insulated gates don't draw current, but the control voltage should be +4 V for maximum gain, and complete cutoff is obtained at −3 V. The first gate is returned to ground. There's no chance of rectified current causing the whole receiver to block when the rf stage conducts (a common cause of receiver overload).

A triode FET can replace the 6J7 mixer. The high feedback capacitance is a minor problem so long as the input and output circuits are tuned to widely different frequencies. The conversion gain is a function of the available oscillator drive power. With enough injection the mixer gain could be ten times that of the original. To replace the 6C5 oscillator a low-transconductance triode FET is needed, as discussed below.

I went to double conversion to obtain better selectivity. If 455-kHz filters were to have been used, I would have moved the first i-f to 930 kHz or higher in order to avoid two intermediate frequencies in harmonic relationship. Of course the second oscillator would have been put on the high side. (Think where the image would be if it were on the low side.) However, I had some 250-kHz mechanical filters on hand. The point of this conversion was, after all, to use the six-band, twenty-four-coil tuning appara-



tus of the original receiver with as little modifications as possible.

Fig. 1 is a block diagram of the complete receiver, including auxiliary features (to be covered in another article). Everything fits inside easily. **Fig. 2** is a schematic of the main rf and i-f circuits.

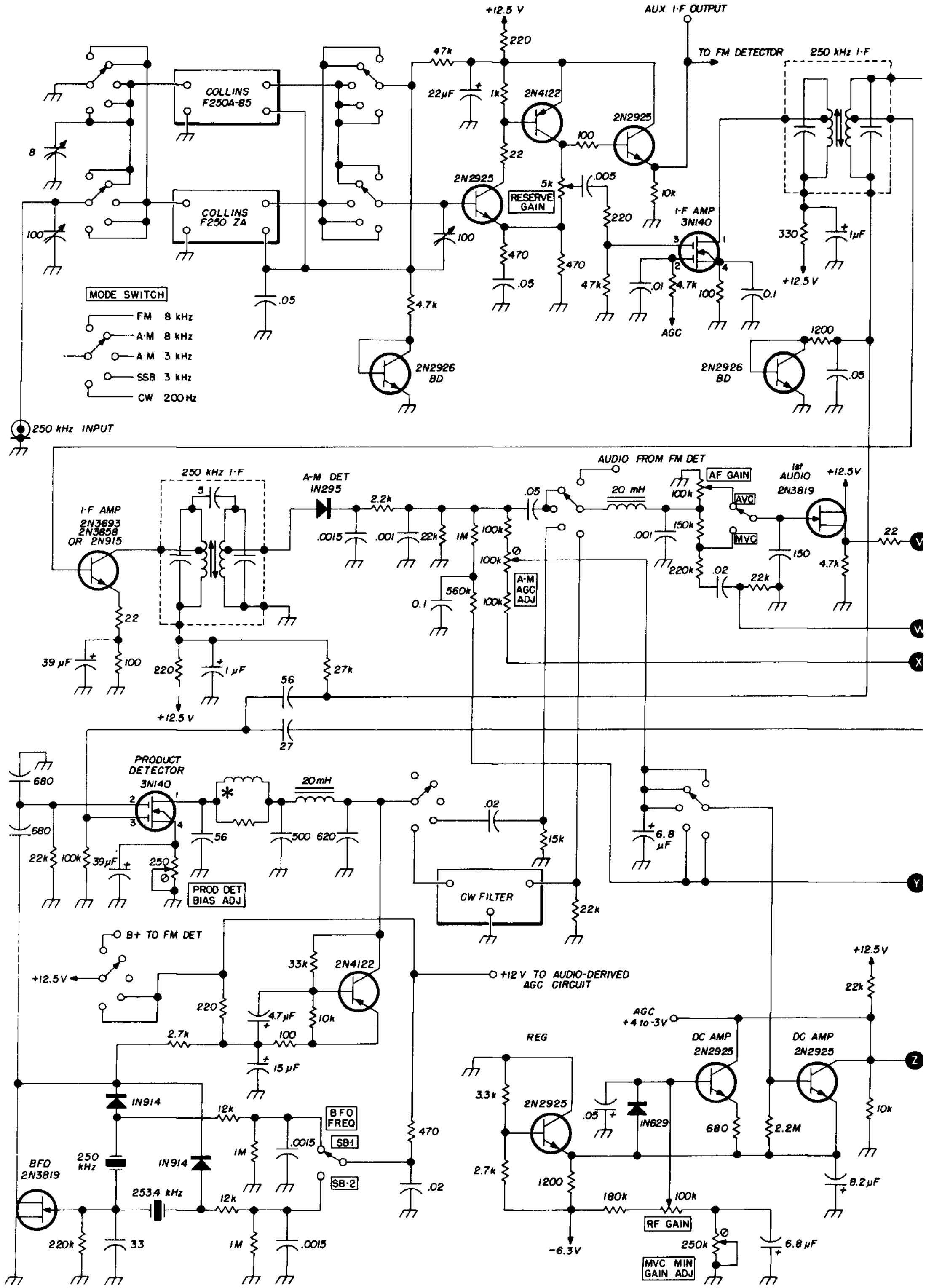
input circuit

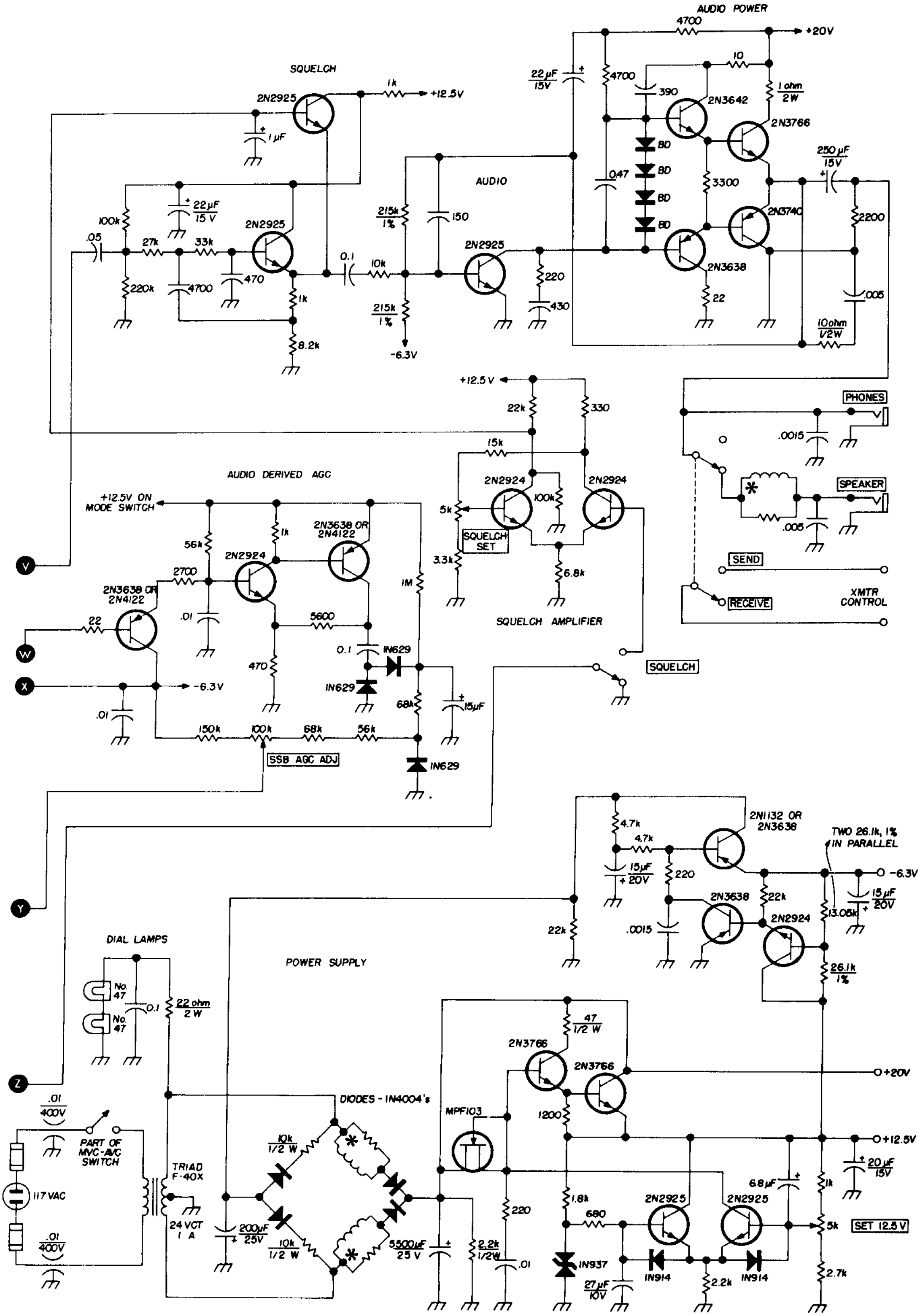
The antenna input starts out with a low-pass filter, with nominal cutoff at about 30 MHz. Aside from obvious advantages if you are near strong TV stations (I am), there are also less "birdies" when the receiver is used with a vhf converter. Because of the input coupling method in this model receiver, the filter is especially important.

There is a protective diode clamp on the antenna coil. 6BA6 rf amplifiers will stand several hundred volts swing without damage. If the grid return resistance is high, however, the diode is essential because the insulated gate FET is rated for only a few tens of volts peak. A type 1N914 of any manufacture will have low loss at 12 volts of reverse bias. The added capacitance is about 1 pF.

The change of drain (plate) current in the first rf stage is used to indicate signal strength. Because of the slope of the 3N140 control characteristic, the meter scale is non-linear in dB, even with the diode in the circuit. Other approaches, including a separate signal indicator channel, were less satisfactory.*

* The use of an s-meter as a quantitative measuring device for received signals is open to question, but an s-meter is fine as a tuning indicator. **Editor**


+12.5 V
AUX I-F OUTPUT
TO FM DETECTOR
250 kHz I-F
COLLINS F250A-85
COLLINS F250 ZA
MODE SWITCH
FM 8 kHz
A-M 8 kHz
A-M 3 kHz
SSB 3 kHz
CW 200 Hz
250 kHz INPUT
2N4122
2N2925
2N2926 BD
RESERVE GAIN
I-F AMP 3N140
AGC
I-F AMP 2N3693 2N3858 OR 2N915
A-M DET IN295
AUDIO FROM FM DET
AF GAIN
AVC
MVC
1st AUDIO 2N3819
A-M AGC ADJ
PRODUCT DETECTOR 3N140
PROD DET BIAS ADJ
CW FILTER
B+ TO FM DET
+12 V TO AUDIO-DERIVED AGC CIRCUIT
AGC +4 to -3V
REG
DC AMP 2N2925
IN914
IN629
BFO FREQ
SB-1
SB-2
BFO 2N3819
250 kHz
253.4 kHz
RF GAIN
MVC MIN GAIN ADJ
-6.3V


AUDIO POWER
4700
+20V
10
22 µF 15V
4700
390
2N3642
1 ohm 2W
2N3766
BD
BD
BD
BD
0.47
3300
250 µF 15V
2N3740
2200
2N3638
22
.005
10ohm 1/2W
SQUELCH
2N2925
1k
+12.5V
1 µF
22 µF 15 V
100k
2N2925
.05
27k
33k
470
220k
4700
1k
8.2k
0.1
10k
215k 1%
150
AUDIO
2N2925
220
430
215k 1%
-6.3V
+12.5 V
22k
330
15k
2N2924
2N2924
100k
5k
SQUELCH SET
3.3k
6.8k
SQUELCH AMPLIFIER
PHONES
.0015
SPEAKER
*
.005
SEND
RECEIVE
XMTR CONTROL
AUDIO DERIVED AGC
+12.5V ON MODE SWITCH
56k
1k
2N3638 OR 2N4122
2N2924
2700
2N3638 OR 2N4122
.01
22
1M
5600
0.1
IN629
IN629
470
-6.3V
68k
15µF
.01
150k
100k
68k
56k
SSB AGC ADJ
IN629
SQUELCH
V
W
X
Y
Z
4.7k
4.7k
2N1132 OR 2N3638
TWO 26.1k, 1% IN PARALLEL
-6.3V
15µF 20V
220
2N3638
22k
15 µF 20V
13.05k
2N2924
22k
.0015
26.1k 1%
DIAL LAMPS
No. 47
No. 47
0.1
22 ohm 2 W
POWER SUPPLY
47 1/2 W
2N3766
2N3766
+20V
1200
MPF103
+12.5V
20 µF 15V
DIODES - IN4004's
.01 400V
PART OF MVC-AVC SWITCH
117 VAC
TRIAD F-40X
24 VCT 1 A
.01 400V
10k 1/2 W
10k 1/2 W
200µF 25V
5500µF 25 V
2.2k 1/2W
220
.01
1.8k
680
2N2925
2N2925
6.8µF
1k
5k
SET 12.5V
IN937
IN914
IN914
27µF 10V
2.2k
2.7k

local oscillator

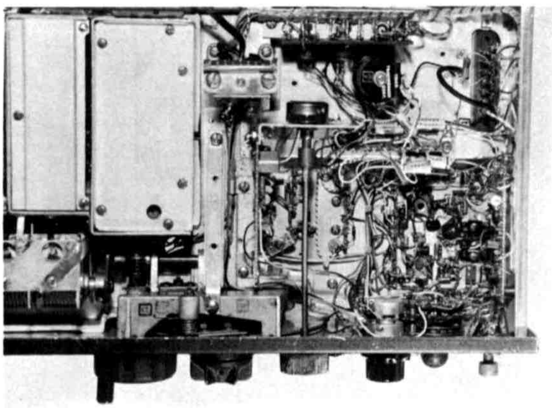
The local oscillator FET socket is supported by its leads, which pass through the keyway from the octal socket pins. The 6C5 ran at around 90 volts, with a starting g_m of less than 2 millimhos. I tried various adaptations, and found that "hot" FET's (those having high I_{dss}* and a high zero-bias g_m) were noisy and unstable, especially at about 15 MHz.

Good results as a local oscillator were obtained with the Motorola 3N126 and some MPF103's. The 3N126 was used with the second gate (substrate and case) hooked to source. If a MOSFET is used, something like a 1N34 would be needed to generate "grid-leak" bias.[1]

In the original oscillator the grid leak was 100k ohms. In the course of chasing troubles I changed the value to 47k, but 100k probably would be satisfactory. The 15k mixer cathode resistor, also inside that box, probably would be satisfactory without change, although mixer FET's having very high I_{dss} would be more suitable in that case. The 3.3k resistor I used was based on experience with the breadboard model of the second converter. The 33- (or 47-ohm) resistor in series with the gate lead of the oscillator can be put in between the octal socket (pin 5) and the transistor socket.

mixer

The mixer proved to be rather critical as to the FET that was used. For type MPF103 the I_{dss} range is 1 to 5 mA; for the 2N3819 it is 2 to 20 mA. I found that anything between 1 and 4 mA worked in my receiver, but the higher-current units gave a bit more gain. Still higher I_{dss} units had low gain and were noisy in this socket. Other places in the receiver are not so finicky in this respect. I'm not worrying about spares: if the FET's last as long as 6SK7's, I'll need another one in 1990.



In a double-conversion receiver, there should be as much selectivity at the first i-f as other considerations permit, and as little gain between the first and second mixer as possible without compromising noise figure (but in no case less than unity gain, or the first stage will overload before the second). A few calculations show that three very high-Q tuned circuits will reject the nearest spurious response (789 kHz), but four tuned circuits of moderate Q will do even better.

I could have used two of the original cans, but I rewound a couple of miniature J-trans to save space. (For details on how to adjust the coupling using a Q-meter, see reference 2.) The "tee" of capacitors between the cans is used in setting up the gain, as the coupling factor between pairs is much less than critical. The shunt capacitance is that of the coax running from the rf board to the second converter module.

second conversion and detector stages

The second mixer and oscillator are built on a small plate that fits in place of the original output transformer. The values were juggled to get best operation with the FET's and crystal I had. With my setup, the oscillator drain swing was about 9 V p-p, and the mixer developed no rectified bias across the 220k resistor.

Strong second oscillator harmonics were found at the mixer output and on the B-supply lead beyond the feed-through capacitor. The additional rf choke and capacitor fixed the power-lead leakage, while the natural attenuation of the mechanical filter is enough protection for the output harmonics,

* I_{dss} is a common classification parameter, measured by shorting the gate to source and applying 5 to 10 V to the drain. Thus by definition, I_{dss} is the drain-to-source current when the gate-to-source voltage is zero.

table 1. comparison of tube and semiconductor characteristics.

Type		V_B (volts)	I_B (mA)	g_M (millimho)	C_{FBK} (pF)	R_{in} at 18 MHz	R_{out} (ohms)	R_{eq} (noise) (ohms)
6C5	tube	80	6	2	2	50k	10k	—
6J7	tube	100	2	1	0.005	60k	800k	10k
6K7	tube	100	9.5	1.6	0.005	50k	150k	15k
2N2084	padt	12	1	34	2.0	800	40k	200
3N126	jfet	12	2	1.4	0.5	25k	30k	—
3N128	mosfet	15	2	4	0.2	100k	80k	—
3N140	mosfet	12	10	8	0.02	100k	12k	$<$1k
MPF103	jfet	12	2	3	2	100k	50k	$<$2k
40481	npn	7.5	2	64	0.2	1500	50k	200

if the filter input terminals, etc., are well shielded.

Second mixer output is parallel-fed via a choke and blocking capacitor to the filter switch. Either the 3-kHz or the 8-kHz bandpass filter may be selected, the other one having its terminals shorted.

The ouput section of the switch (8 pole, 5 position) goes to a feedback amplifier that has a high-impedance input and a gain of ten, followed by a potentiometer that permits the background noise level to be adjusted without modifying the agc action. The range of the control is 23 dB.

The stage following the gain control is the third 3N140. It is coupled by an auto radio i-f can, padded down from 262 to 250 kHz with additional 22-pF capacitors, to the product detector and to an i-f power amplifier that drives the diode detector. The two i-f cans shape the top of the passband in the 8-kHz bandwidth a-m position to get the best weak-signal a-m reception. (This seems to want a rounded top, symmetrical, down about 6 to 10 dB at 3-kHz off center.)

The front-panel switch selects five combinations of bandwidth and detection, as per **table 2.** The agc source for a-m is the diode detector, but for sideband and CW reception agc is derived from rectified audio, so that the effect of audio selectivity is included.

mode-selection circuit

The mode switch is shown for 8-kHz bandwidth a-m reception, with agc in operation. The second-gate control voltage on the three 3N140's varies from plus three or four volts with no signal to negative two or three on very strong signals. The loop gain is high, giving a dynamic range of 40 dB. The first volt of bias change makes very little change in receiver gain, but the squelch operates reliably at that point.

agc/mvc

In the gain control circuit the two dc amplifiers, one to handle agc and one for manual control, are hooked to a transistor that supplies −4 V to the emitter returns and which, incidentally, provides temperature compensation—a sort of three-transistor differential amplifier.

A negative voltage supply makes the am-

table 2. Mode switch setup.

function	FM-8	AM-8	mode AM-3	SSB	CW	REMARKS
filter (kHz)	8.5	8.5	3	3	3	4 – poles
200-Hz audio filter	no	no	no	no	yes	1 – pole
agc	car	car	car	af	af	1 – pole
B+ to SSB and bfo	off	off	off	on	on	
audio from	fm	a-m	a-m	prod	prod	1 – pole
B+ to fm det	on	off	off	off	off	1 – pole for B+

notes:

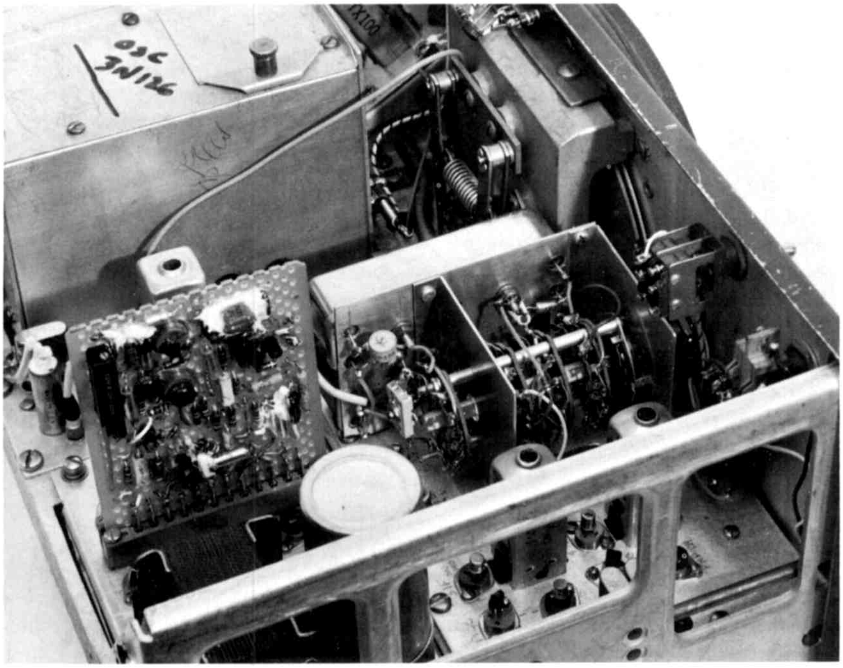
1. SB1-SB2 (SB1 is lower 40-75, upper, 20).
2. Agc-off-mvc functions similar to original.

plifier practical and is a handy way to make dc level changes without extra stages. Also the audio stage operating point is centered automatically.

The audio-derived agc voltage is from a little three-transistor feed-back amplifier (with automatic temperature compensation; again, this is from the pnp follower driving the npn stage) and a voltage-doubling rectifier. The attack time is fast for small changes, but the maximum effect from a single pulse is limited by a 0.1-microfarad capacitor that couples the amplifier to the rectifier, while the rectifier has fifteen microfarads to charge up.

For 1-kHz CW tones, this means that 1 Hz makes a maximum difference in charge of the 15-μF capacitor of 10 V times 0.1 divided by 15, or 0.067 V. However, the gain of the dc amplifier after that is considerable, and gain changes of 20 to 40 dB/millisecond can be obtained. If action were much slower, there would be worse thumping at the beginning of a sideband transmission. If the action is too fast, there may be a stability problem or there may be a tendency for the receiver to block on a burst of QRM or ignition noise.

For fm reception, a Sprague integrated circuit is used, type ULN-2111A. This has a bunch of emitter-coupled clipping amplifiers, a rather fancy six-transistor phase detector, and various diode bias networks. All that is required outside is a single-tuned circuit. (A typical transistor radio i-f can is suitable; just use the high impedance winding, padded to frequency with external capacitors and some by-pass capacitors.) The output in my receiver swung three volts peak-to-peak for a signal moving across the 8-kHz filter. When the signal was injected after the filter, an s-shaped curve about 18-kHz wide was obtained, with an amplitude of more than five volts peak-to-peak.



Chassis view of the revamped BC-348 shows the new mode switch and two mechanical filters. The power supply and audio circuits occupy the old dynamotor space; the plug-in board contains voltage regulating and low-level audio circuits.

For manual gain control (I happen to like to dive for the same volume control knob whether agc is on or not), the agc circuits are still operative to prevent blocking on unexpectedly strong signals. The audio gain is high enough so that normal operation is in a region where the agc threshold is not exceeded. Because of the highly nonlinear control function (dB versus control volts), tapered sections are used for both audio and manual rf gain.

diodes. However, 1N914's could be used in place of the 1N457's as well.

The 2N2925's and 2N2924's were actually the green- or yellow-coded 2N2926's, while the brown- and red-coded 2926's were used for bias compensation diodes, labeled "BD." You get 25 assorted for $7.25.** You'll be able to use 20 if you're crazy enough to copy this receiver.

There are also two 2N3638's, three 2N4122's (two in the calibrator, which is

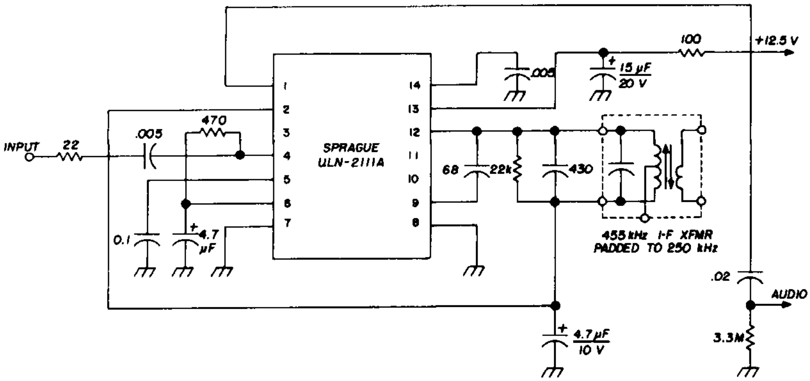


fig. 3. Integrated-circuit fm detector using a Sprague ULN-2111A. This circuit provides 3 volts p-p output with 1 millivolt input.

I found that a dual 100k audio taper unit was satisfactory; the value could be as high as 250k and as low as 50k with a few minor adjustments. A stereo amplifier control should be adequate. The **reserve gain** pot in the 250-kHz i-f amplifier could have either a linear or audio taper such as used in transistor radios.

The rest of the controls (for agc thresholds, product detector bias, manual volume control "off" bias,* and the power supply voltage setting) can be any handy trim pot, either surplus from computer cards, or small trimmers such as the IRC 2CI or Mallory MTC.

bfo circuit

High-speed computer diodes such as the 1N903 or 1N914 are used for the input voltage clamp and the bfo crystal switch; I used 1N457's or 1N629's for other spots, as I had them, and I felt that there was less chance of generating harmonics with the "slow"

mounted on the rf panel near the input) and six pnp's, which could be either type. The first type costs fifty cents; the second eighty. (The whole count is 50 transistors and FET's, 16 various diodes, and one integrated circuit.)

audio section

After the mode of reception is selected (with or without narrowband CW filter), the signal goes, via the audio gain pot in the agc position, to a source-follower first audio stage, an active low-pass filter, then to a complementary class-B output stage. The squelch is in the active filter section. The circuit is

* Minimum gain setting must be at minimum gain bias.

** Available from Allied Radio Corporation, 100 N. Western Avenue, Chicago, Illinois 60680. Order stock number 49F3-2N2926. Kit of 25 is $7.25 plus postage.

similar to one previously described.[3]

The power regulator, power transformer (Triad F-40X 26 V, 1 A center-tapped), squelch control, and all except the first audio are mounted in the space vacated by the dynamotor. The high-power transistors are on the subchassis (insulated by mica washers), and the low-power circuits are on a plug-in card (very easy to pull out and change).

selectivity filter

The CW filter is switched into the sideband channel for more selectivity when needed. The values seem reasonably satisfactory in practice, although the filter performance has not been measured in place in the receiver. Any other narrowband audio filter of ten or twenty thousand ohms impedance should do as well. Obviously, an extra amplifier stage could be put in the chain if the filter on hand was of lower impedance, as the method of switching makes this practical. The audio agc is based on what passes through the filter, so it can be used even under quite rough conditions. Because the squelch can also be used in this mode, I have been able to find a fairly weak (though **stable**) signal and park on him with the receiver on squelch, waiting to get a chance to work him.

crystal calibrator

The crystal calibrator is almost a necessity in a wide-range unit such as the BC-348. It appears that there would be some benefit in making a more complex calibrator, such as a 200-50-10 kHz type. Higher output than I have is desirable when running some converters, but I haven't done anything about it as yet.

in retrospect

The main problems with transistors almost always stem from overloading and temperature variation. Resolving the overload problem takes effort: care to see that the last i-f amplifier, for instance, is capable of driving the second detector hard enough, so that the second detector can generate enough agc voltage, so that it won't be driven too hard—like a tight servo system. (The more I read that the more I wonder if the English language is up to today's problems.)

What all this means is that the last i-f stage should be designed for best power output, not best gain. Also, it means that agc can't be used on the last i-f stage. (Auto radios with **one** i-f stage have some rather fancy solutions to this problem.)

Another device to reduce overload is feedback. At "low" frequencies (under a couple MHz for today's transistors), feedback will hold gain and bias constant despite temperature variation. Distortion will also be kept low.

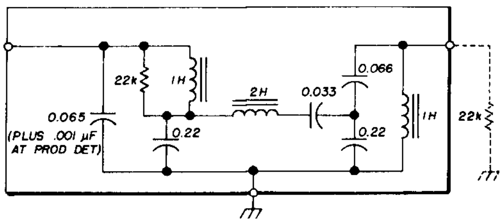


fig. 4. Audio CW filter with nominal 700-Hz center frequency and 160-Hz bandwidth. Inductors are Triad epoxy-molded toroids with Q of 50 to 80 at 1000 Hz. Capacitors are mylar insulated.

In many places where dc levels are critical, differential amplifiers are used. The drift with temperature of one transistor is balanced almost exactly by another of the same type. In other circuits, a diode-connected transistor (or any diode) is used as part of the dc bias circuit, so that temperature effects of the transistor are pretty well cancelled by the diode as it conducts a few mA in the forward direction. In class B or AB circuits this is particularly important. For best results, the bias diodes must be placed physically so that they are at about the same temperature as the power transistors.

references

1. G. D. Hanchett, "Insulated Gate Field Effect Transistors in Oscillator Circuits," *RCA publication ST-3520.* (Available from P. O. Box 53, Harrison, N. J.)
2. F. Langford-Smith, *"Radiotron Designer's Handbook,"* 1952 edition, page 1030.
3. H. H. Cross, W1OOP, "An Obsolete Mobile Receiver," *QST*, November, 1967, p. 11, and December, 1967, p. 31.

ham radio